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EDITORIAL

The Radiolocation Convention

THE Convention which was held at the Institution of Electrical Engineers in London from 26th to 29th March must be regarded as one of the milestones in the history of radio. The attendance at the inaugural meeting showed that the theatre of the Institution is too small for such occasions; overflow meetings had to be held in two other rooms. It will be recalled that when Sir Edward Appleton gave the Kelvin Lecture last year on "The Scientific Principles of Radiolocation" a large number of people failed to get into the theatre and the lecture had to be repeated. We note that the forthcoming Kelvin Lecture by Professor Oliphant on Nuclear Physics has been transferred from the Institution to the Kingsway Hall—a wise step. As a result of the war the number of people interested in the scientific development of radio and other branches of applied science has greatly increased and there is a very wide-spread desire to learn the details of the various developments that have taken place during the war, but which have necessarily been kept secret. On 8th April the Physical Society and the Royal Meteorological Society held a joint conference from 2.15 to 7.30 when nine papers were read on various aspects of "Meteorological Factors in Radio Wave Propagation." A convention was held in New York on 23rd to 26th January by the American Institute of Radio Engineers when 87 papers were read. Very brief summaries—sometimes merely the titles—of these are given in the February number of the Proc. I.R.E. with the tantalizing statement that "no papers are available in preprint or reprint form, nor is there any assurance that any of them will be published in the Proceedings, although it is hoped that many of them will appear in their pages." It was interesting to note that Dr. F. B. Llewellyn, the President of the American I.R.E., was able to be present at the London Convention and gave an address at the opening meeting. Speaking of the close co-operation between the two countries, he said that it was possible that radar was the best example of co-operative technical effort that the world had ever known. The other two speakers at the inaugural meeting were Mr. Wilmot, the Minister of Supply and Aircraft Production, who opened the proceedings, and Sir Robert Watson-Watt who gave a general survey of the evolution of radiolocation, a task for which he was eminently suited as he has been engaged in it, in some form or other, for over twenty years.

The first session of the Convention proper was devoted to Aerials and Waveguides. Each session opened with one or two introductory, survey, or integrating papers, and at this first session one on aerials was read by J. A. Ratcliffe and another on waveguides by Dr. Pryce. Then followed seven papers as follows:—Ground Radiolocation Aerials, by D. Taylor; Slot Aerials, by H. G. Booker; Slotted Linear Arrays, by D. W.

Fry; Cheese Aerials, by O. Böhm; Waveguide Matching Technique, by E. Wild; Phase Correction of Horn Radiators, by N. M. Rust; and Production Testing of Waveguides, by L. W. Brown. Dr. Pryce's paper has a useful bibliography of 17 forthcoming papers on waveguides based on service reports and an indication in each case as to where they are likely to be published. Ratcliffe's paper has reference numbers up to 31, but unfortunately the bibliography is missing from the paper; it will doubtless appear later. Perhaps the most novel of all the subjects dealt with in the above papers was the replacement of the ordinary aerial by a slot of the same size cut in a metal sheet. This is an application to radio of Babinet's principle in optics which deals with the equivalence of an illuminated slit in an opaque screen and an opaque rod in a beam of light. If the slot is energized by connecting a generator between the two midpoints of the opposite sides of the vertical slot, the electric field will obviously be horizontally polarized, while the magnetic field will lie in vertical planes. Compared with a vertical dipole the directions of the electric and magnetic fields are thus interchanged.

Propagation

For the second session one had to choose between Propagation and Cathode Ray Tubes as these two sessions were held simultaneously. In the Propagation section there were two preliminary survey papers, one by Dr. Booker on Elements of Radio Meteorology and the other by E. C. S. Megaw on Experimental Studies of the Propagation of Very Short Radio Waves. Here, again, there were seven contributions, viz., Measurements of Refractive Index Gradient by F. L. Westwater; Attenuation of Centimetre Waves and Echoes Resulting from Atmospheric Phenomena, by J. W. Ryde; Transmission Conditions over Land and Sea, by Dr. Smith-Rose; Microwave Propagation, by B. S. Starnecki; Theoretical Estimation of Field Strength, by J. M. C. Scott; the Importance of Theory, by T. L. Eckersley, and finally, after a discussion of the foregoing, a paper on Extra-tropospheric Influences on U.S.W. Communication, by Sir Edward Appleton.

Perhaps the outstanding feature of this session was the stress laid on the propagation of microwaves over distances far beyond the visible range by radio ducts or atmospheric waveguides close to the surface of the sea. In this connection it is interesting to note that one of the papers read at the American convention was entitled "Three and Nine Centimetre Propagation Measurements in Low-level Ocean Ducts." Another interesting phenomenon was the echo from moderate rain some tens of kilometres distant, which might be comparable with that from an aircraft at the same distance. Such effects increase rapidly with increasing frequency.

In the cathode-ray tube session two survey papers were read giving a general survey of the wartime developments, one by J. G. Bartlett, D. S. Watson and G. Bradfield, and the other by L. C. Jesty, H. Moss and R. Puleston. These were followed by three papers on screens, the first by G. S. J. Garlick, S. T. Henderson and R. Puleston, the second by R. G. Hopkinson and the third, which dealt with the Skiatron or dark trace tube, by P. G. R. King. In the evening a lecture on "Precision Radar" was delivered by W. A. S. Butement who gave a very complete account of the development of the subject, and of the various alternative methods with their advantages and disadvantages.

The morning of the 28th March was devoted to valves, the survey paper on which was by Dr. J. H. E. Griffiths. This was followed by seven papers as follows: Early Work on the Cavity Magnetron by J. T. Randall and H. A. H. Boot; High Power Pulsed Magnetron, by W. B. Willshaw and L. Rushforth; Velocity Modulation Valves, by L. F. Broadway; The Crystal Valve, by B. Bleaney; Gas Discharge Switches, by A. H. Cooke; Triodes for Very Short Waves, by J. Bell; and Gas-filled Triodes for Radar Modulator Service, by H. de B. Knight. The papers on the magnetron will probably be regarded by most people as the high-water mark of the Convention and detailed papers dealing with the subject will be eagerly awaited; as Griffiths says, "This work was outstandingly successful, and resulted in what was probably the greatest single contribution to radar." It is a matter of great satisfaction that this idea, which originated with Randall and Boot at Birmingham University, was developed and brought to such outstanding fruition by the joint efforts of the University, the G.E. Co., the M.O. Co. and the B.T.H. Co. The paper by Bleaney is interesting because it describes how the old crystal detector with its cat's-whisker has been re-instated

and it is interesting to read that "no vacuum tube which can rival its sensitivity has been developed for centimetre wavelengths". Radio measurements was the subject of the following session, and a survey paper was read by C. W. Oatley entitled "Ultra High Frequency Measurements," which was followed by seven papers as follows: Some Monitors of Transmitted Power at Centimetre Wavelengths, by L. B. Turner; Gold Leaf Electroscope and Enthrakometer by J. Collard; Signal Generators and Noise Measurement, by B. Bleaney; Cable Measurements, by L. Essen; Resonance Methods of Dielectric Measurement at Centimetre Wavelengths, by Willis Jackson; Radiation Patterns from Directive Antennae, by J. Dyson and B. A. C. Tucker, and the Study of Centimetre Wave Propagation, by H. Archer-Thomson and E. M. Hickin. The paper by Prof. Willis Jackson has exactly the same title as one published by him and a number of associates in the January number of Part III of the Institution Journal, and presumably covers much the same ground. Measurements at these extremely high frequencies bristle with difficulties and, when published, these papers will give a very good review of the methods developed to overcome these difficulties in a very wide range of measurements.

An evening lecture on "Shipborne Radar," by A. W. Ross described the difficulties due to the large accumulation of radio gear on board ship. It was surprising to learn that some ships have more than 30 radar sets in addition to an equally large number of other sets for radio communication. The complexity of the problems involved in avoiding interference, in frequency allocation and screening can easily be imagined. These matters were discussed in the lecture. It was stated that the 3-metre transmitter first fitted in 1940, which had a peak pulse power of 1,000 kilowatts, was the most powerful radar transmitter in service during the war. Another speaker mentioned peak powers of 3,000 kilowatts. Even although such a pulse may only last a few microseconds it can easily prove a very disagreeable neighbour unless special precautions are taken.

The morning session on 29th March was devoted to Transmitters and Receivers, with survey papers by O. L. Ratsey on the former and W. B. Lewis on the latter. Four papers on the development of various types of transmitter were read by R. V. Whelpton and J. M. Dodds, T. S. England, K. J. R. Wilkinson, D. F. Gibbs and B. W. Lythall, and four on receivers by H. Dahl, W. L. Watton, F. L. Humber, R. S. Paulden and M. K. Taylor, G. G. Macfarlane and J. R. Whitehead. The transmitters described cover a very wide field of research and development and a list is given of thirteen papers which are being prepared with a view to publication in the Journal of the I.E.E. The paper on the super-regenerative receiver takes one's mind back twenty years or more when such receivers were very much discussed.

Pulse Circuits

The subject for the afternoon session was "Circuit Techniques and Radiolocation" with an introductory paper by F. C. Williams, followed by seven papers as follows: Blocking Oscillators, by R. Benjamin; Low Power Pulse Transformers by N. F. Moody; Delay Networks in Pulse Formation, by E. L. C. White; Very Short Pulses, by D. C. Espley; Light Weight Radar, by H. R. Whitfield; The Velodyne, by F. C. Williams and A. M. Uttley; and Automatic Strobes, by F. C. Williams and F. J. U. Ritson. The introductory paper of 21 pages discusses many circuit problems in a very interesting manner and gives a bibliography of 38 references, several of which are described as I.E.E. convention papers and lectures, although they do not appear on the programme of this convention.

This session was followed by a lecture on "Radar Navigation" given by R. A. Smith.

An additional meeting was held on 3rd April when five papers on naval gunnery radar were read by J. F. Coales, J. C. Calpine and D. S. Watson, R. V. Alred, C. A. Laws, H. W. Pout and H. A. Prime.

The publication of the papers read at the Convention and of the many other papers referred to as being prepared for publication will be eagerly awaited; it will fortunately take a considerable time even with all the resources of the Institution journal and the technical press to publish such a great number of papers, many of which are of considerable length. A steady output spread over many months will facilitate the processes of digestion. The Institution of Electrical Engineers and all concerned in the organization of the Convention are to be congratulated on the highly efficient way in which it was carried out. G. W. O. H.

CAVITY-RESONATOR WAVEMETERS*

Simple Types of Wide Frequency-Range

By L. Essen, B.Sc., Ph.D., A.M.I.E.E.

(*Communication from the National Physical Laboratory*)

SUMMARY.—The description is given of four simply constructed cavity-resonator wavemeters covering the frequency ranges 10,000 Mc/s–4,000 Mc/s, 5,600 Mc/s–2,000 Mc/s, 2,700 Mc/s–1,000 Mc/s and 1,000 Mc/s–200 Mc/s. The mode of resonance employed is the hybrid between the cylindrical TM_{010} mode and the coaxial TM_{00p} mode, and the frequency variation is effected by the axial movement of a plunger attached to a micrometer head. For the first three wavemeters the plunger is of a non-contact design, thus obviating the necessity for a good electrical contact which has hitherto caused considerable manufacturing difficulties. The setting accuracy of the instruments is shown to be better than 1 part in 10^4 of frequency throughout the greater part of the range.

Introduction

THE increasing application of electromagnetic oscillations of centimetric and decimetric wavelengths in recent years has created a need, particularly in laboratory work, for simple portable wavemeters to cover the complete frequency range between 200 Mc/s and 10,000 Mc/s. For most applications an accuracy of 1 part in 10^4 is adequate, and the range covered by each instrument should be as large as is consistent with this accuracy.

It is well known that electromagnetic waves are freely transmitted along hollow metal tubes, or waveguides, the dimensions of which exceed certain critical values depending on the wavelength of the oscillations. If a length of waveguide is closed by metal plates at each end, reflections of the wave occur and a system of standing waves is established. The closed waveguide constitutes an electrical resonator which, if coupled to a source of oscillations and provided with a means of detecting the oscillations, forms a very simple and convenient wavemeter. It can, however, resonate in many different modes having different frequencies, and for wavemeter applications it must be designed to have the greatest possible frequency separation between neighbouring resonances. Such cavity-resonator wavemeters have been widely used during the last few years.

For the purpose of designation, the modes are divided into two main types, the transverse electric (TE or H) modes, in which there is a component of magnetic field, but not of electric field, in the direction of propagation and the transverse magnetic (TM or E) modes in which the reverse is true. In the case of a circular guide the modes are uniquely defined by the addition of three suffixes which give the number of periodic variations of electric field in the angular, radial, and axial directions respectively. The value of the first suffix corresponds to the number of full period variations of electric field in a rotation of 360°, the value of the second to the number of half-period variations across a diameter and the value of the third to the number of half-period variations in the length of the guide. The configurations of the fields in the different modes are shown diagrammatically in a number of text-books.

The coaxial resonator, in which there is an inner conductor in the tube, may be regarded as a special case, its mode designation being TM_{00p} or TE_{00p} since there is no component of magnetic or electric field in the direction of propagation. The value of p is the number of half-waves along the coaxial line.

Barrow and Mieher† described a form of cylindrical waveguide resonator with which a frequency range of more than 2 to 1 could be covered by the axial movement of a plunger, the mode of resonance being the hybrid between the cylindrical mode TM_{010} and the coaxial mode TM_{00p}. A wavemeter using this mode was designed by Dr. Sayers at Birmingham University, but difficulty was experienced in obtaining a sufficiently reliable electrical contact between the plunger and the wall of the resonator. Moreover, in order to reduce the wear of the plunger, it was heavily chromium-plated. This resulted in a relatively high electrical resistance with

* MS. accepted by the Editor, October 1945.

† W. L. Barrow and W. W. Mieher. "Natural Oscillations of Electrical Cavity Resonators," *Proc. Instn Radio Engn*, 1940, Vol. 28, pp. 184–191.

the result that the Q-factor (π/Log. Decrement) of the wavemeter was reduced as the plunger was inserted. These difficulties have been overcome by the use of a non-contact plunger in three of the wavemeters described in this paper. These cover between them the frequency range 1,000 Mc/s to 10,000 Mc/s. In the wavemeter of the lowest frequency range, 200 Mc/s to 1,000 Mc/s, it was found that a contact type plunger was necessary to preserve a convenient physical design of instrument.

1. General Features of the Design using a Non-Contact Plunger

The general design of the wavemeters is illustrated in Fig. 1, and the photograph shows the three non-contact models. In the mode of oscillation used there is a flow of current between the plunger and the walls of the resonator, and there must therefore be a low impedance path between them. The input impedance at C, Fig. 1, is that of a coaxial line, including lengths of very different values of characteristic impedance.

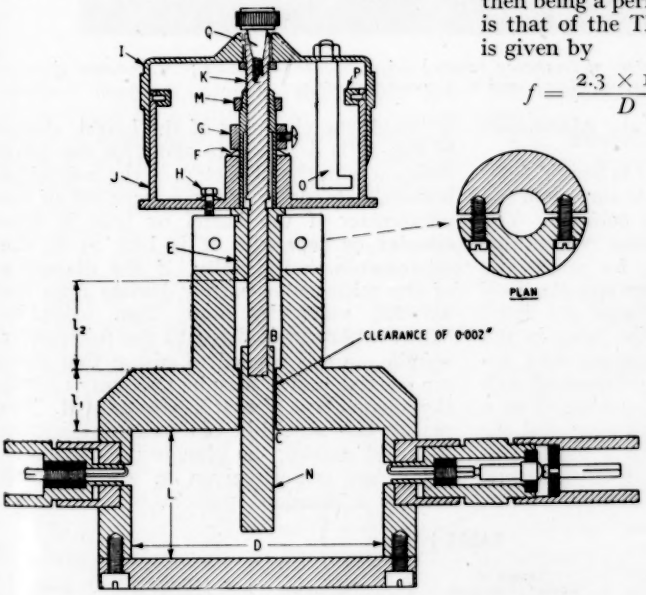


Fig. 1. *This section through a wavemeter shows the general design adopted for types A, B and C. A micrometer-head is used for the adjustment of the plunger N.*

Neglecting the smaller discontinuity at B caused by the different diameters of the plunger and the micrometer spindle and neglecting the resistance of the conductors, the input impedance is

$$Z = jZ_1 \frac{\frac{Z_2}{Z_1}\tan \beta l_2 + \tan \beta l_1}{1 - \frac{Z_2}{Z_1}\tan \beta l_2 + \tan \beta l_1} \qquad (1)$$

in which Z_1 and Z_2 are the characteristic impedances of the lines of lengths l_1 and l_2, respectively, and $\beta = \frac{2\pi}{\lambda}$, λ being the wavelength of the oscillations. With the dimensions used, and a clearance of 0.002 in, the values of Z_1 and Z_2 are of the order of 1 ohm and 50 ohms respectively and the input impedance Z is small unless l_1 or l_2 approaches 0 or $\frac{n\lambda}{2}$. Values ranging between 0.06λ and 0.42λ have been used with satisfactory results. The dimensions of the resonating element of the wavemeter are governed by the upper frequency limit, and the need to avoid the excitation of other modes of oscillation within or near the frequency range of the instrument. The highest frequency is obtained when the plunger is fully withdrawn, the resonator then being a perfect cylinder. The frequency is that of the TM_{010} mode of resonance and is given by

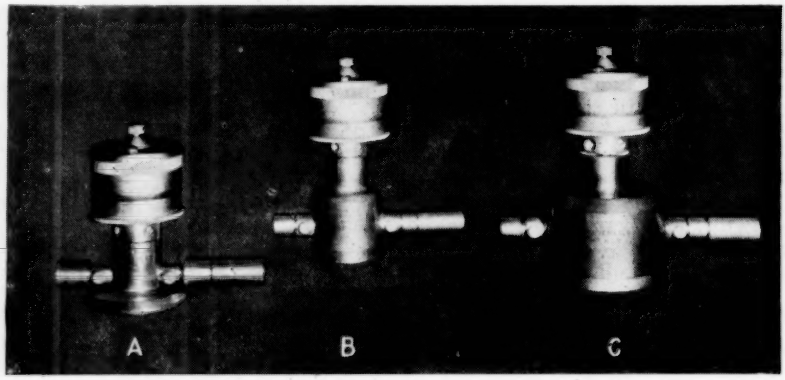
$$f = \frac{2.3 \times 10^{10}}{D} \text{ c/s} \qquad .. \qquad .. \qquad (2)$$

D being the diameter in centimetres.

In order to avoid confusion with the TE_{111} mode of resonance the length of the resonator should be less than the diameter. The TE_{111} mode will then occur at a frequency higher than that of the TM_{010} mode, and there is no mode of resonance at a frequency lower than that of the TM_{010} mode. The frequency determined by the

reading of the wavemeter is thus free from ambiguity.

The movement of the plunger is effected by a micrometer head, a standard commercial model being used because of the difficulty of obtaining articles made to special requirements during wartime conditions. The movement is therefore restricted to 1 in, and for the wavemeters covering the lower frequency ranges it is moreover necessary to extend the spindle of the micrometer as shown in Fig. 1.



Cavity-resonator wavemeters of frequency ranges:—A, 4,000–10,000 Mc/s; B, 2,000–5,600 Mc/s; and C, 1,000–2,700 Mc/s.

The body of the wavemeter is turned from a solid brass rod and the inside surface of the resonator is silver-plated and polished. The plunger is a silver-plated brass rod, except for the smallest wavemeter, for which the lower half of the micrometer spindle itself is silver-plated. Coupling loops are fitted through diametrically opposite holes in the cylindrical walls of the resonator, and are fixed in a suitable position by means of grub screws. The input loop is terminated in a socket to take a G.E.C. Type 3 plug, and the detector loop is in series with a totally enclosed fixed-contact crystal detector, the output from which is taken from a similar socket.

2. Details of the Design using a Non-Contact Plunger

The main electrical requirement is that the plunger should be circular and move coaxially in the hole through the end wall of the resonator. To achieve this, the hole to take the micrometer barrel and that through the wall of the resonator should be bored together without the wavemeter being taken from the lathe. It has been found also that although the clamping of the barrel, shown in Fig. 1, is designed to allow for the small differences in the diameters of the individual barrels, it tends to produce some tilt of the micrometer if the barrel or hole is non-circular or tapered. It is best to fit the micrometers individually, if the diameters of the micrometer barrels deviate from the nominal value by more than 0.0002 in. When a plunger is fitted to the micrometer spindle, care is needed to ensure that these are coaxial, and it may be necessary to turn the plunger true after it has been fitted. The plating also should be of a uniform thickness to avoid making the plunger non-circular.

For the reasons given in the preceding section a standard pattern of micrometer

TABLE I

Designation of Wavemeter	Size of Resonator (cm)		Length of Coaxial Lines (cm)		Frequency Range (Mc/s)	Accuracy of Setting and Reading	Average Value of Q
	D	L	l_1	l_2			
A	1.6	1.47	0.9	1.04	4,000–10,000	1 in 10^4	1,000
B	3.8	2.54	1.9	2.23	2,000–5,600	0.5 in 10^4	1,500
C	7.6	3.80	1.9	2.54	1,000–2,700	0.3 in 10^4	4,000
E	15.2	4.50	Contact plunger with 2 in diameter disc		200–1,000	1 in 10^4 to 10 in 10^4	1,300

head was used. It is convenient in the normal applications of the micrometer to be able to adjust the zero readily, and the drum J (Fig. 1) is therefore held relative to the barrel E only by the friction between the spring washer F and the sleeve G which is pushed down to compress the washer and then locked by a grub screw. This friction fixing is inadequate for the present purpose since a rotation of the drum, on which the vertical scale of the micrometer is inscribed, would alter the calibration of the wavemeter. Such a rotation is therefore prevented by the screw H. In order that this screw should not distort the drum, a light pressure on a spring washer is used. The thimble I of the head is fixed to the spindle K by means of a conical friction clamp Q. This fixing has been found sufficiently secure, although here again a rigid fixing would be preferable.

The tightness of the micrometer screw is adjusted by means of the nut M. A suitable compromise between ease of operation and freedom from backlash is obtained if the thimble continues to rotate for about one turn when it is spun by hand and released.

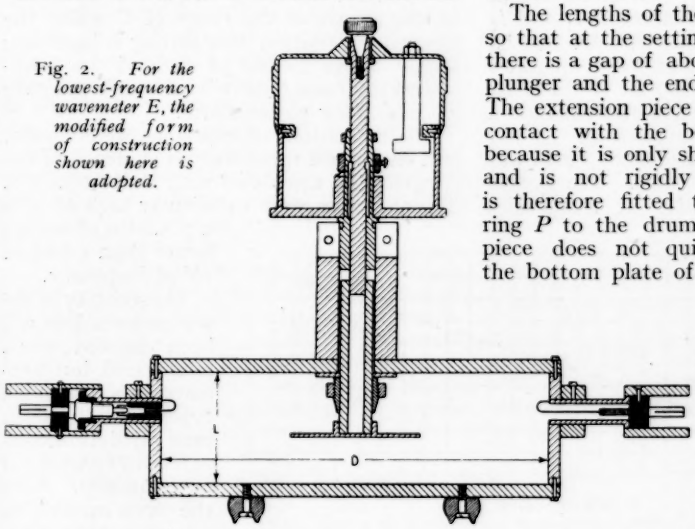


Fig. 2. *For the lowest-frequency wavemeter E, the modified form of construction shown here is adopted.*

The lengths of the plungers are adjusted so that at the setting O on the micrometer there is a gap of about 0.5 mm between the plunger and the end face of the resonator. The extension piece N must not come into contact with the body of the wavemeter because it is only shrunk on to the spindle and is not rigidly fixed. The stop O is therefore fitted to the thimble and the ring P to the drum so that the extension piece does not quite make contact with the bottom plate of the resonator or with the barrel E of the micrometer. A counterweight, not shown in the diagram, is fixed in a diametrically opposite position on the thimble.

The movement of the plunger in wavemeter A is only 1.5 cm and it is unnecessary in this case to extend the micrometer spindle. It is instead silver-plated on one half of its length.

The dimensions of the wavemeters for different frequency ranges are given in Table 1.

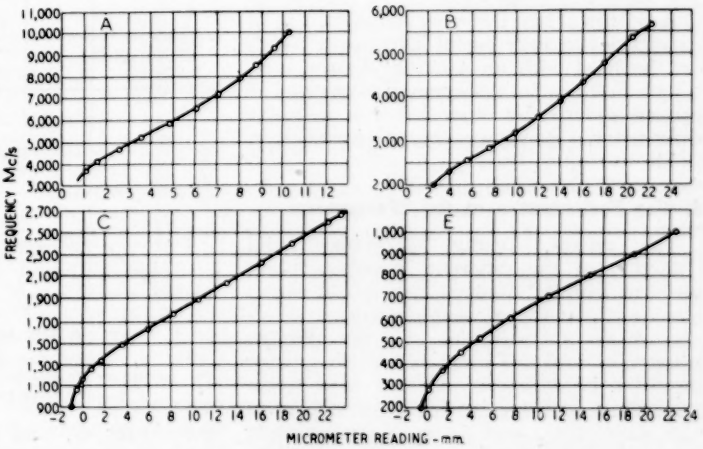


Fig. 3. *Calibration curves of the four wavemeters described.*

3. Details of the Wavemeter of Range 200 Mc/s–1,000 Mc/s.

Experiments were made with a non-contact plunger in a resonator of diameter 11.8 cm and length 4.5 cm, the values of l_1 and l_2 being 2 cm and 3.3 cm respectively. Satisfactory operation was obtained at frequencies between 500 Mc/s and 1,700 Mc/s. In order to obtain lower frequencies of resonance a disc of 2 in diameter was fitted on the end of the plunger, and the frequency range was then 200 Mc/s to 1,000 Mc/s. The instrument was very insensitive at frequencies below 500 Mc/s, however, and it was found to be very much improved when electrical contact was made between the plunger and the walls of the resonator by means of sprung fingers. The model shown in Fig. 2 was therefore designed for this range of frequency. The design is considered to be inferior to that of the other three wavemeters. It is nevertheless a very useful instrument which can be used with an accuracy better than 1 part in 1,000 in a frequency range not at present covered by other forms of resonant wavemeter. Its dimensions are included in Table 1.

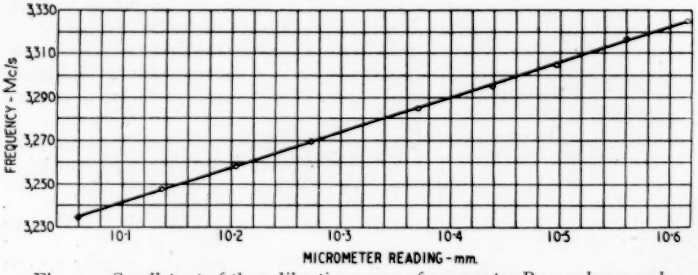


Fig. 4. *Small part of the calibration curve of wavemeter B on a large scale.*

4. The Performance of the Wavemeters

The ranges, average accuracies, and Q-factors of the different wavemeters are given in Table 1; and the calibration curves are shown, on a small scale, in Fig. 3. With this mode of oscillation, as the plunger penetrates into the cylinder, the frequency variation is at first small, it is then larger and fairly uniform for the greater part of the movement, and becomes increasingly large when the gap between the end of the plunger and the end face of the resonator becomes small. The wavemeters are designed so that as far as possible only the middle portion of this range is used. The curves do not show the initial small variation of frequency, but to obtain an overlap between wavemeters C and E it is necessary to use a part of the range of C where the frequency variation with setting is becoming large. Three models of each of the wavemeters A, B and C have been tested and gave a satisfactory performance.

The indications of resonance were smooth and sharp and there was no evidence of the irregularities associated with faulty contacts. The Q-factors were sufficiently high to give a precision of setting better than 1 part in 10^4 of frequency.

The stability of the wavemeters has not been checked over a long period, but measurements made with types B and C at an interval of six months agreed to 1 part in 10^4 of frequency. After the same interval of time the calibration of type A had changed by several parts in 10^4.

Suitable conditions of coupling were obtained with the tip of the loop withdrawn 2 mm from the inside wall of the resonator of the wavemeter A, flush with the wall of wavemeter B, and protruding within the resonators of wavemeters C and E by 1 mm and 3 mm respectively. With these settings the resonant frequencies of wave-

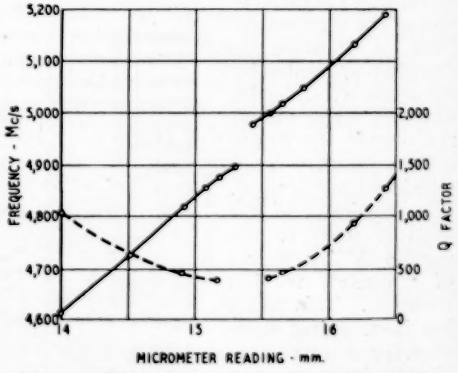


Fig. 5. *These curves show the effect on the resonance frequency and Q-factor of a subsidiary resonance behind the cavity.*

meters B, C and E were not influenced by changes in the input and detector circuits to an appreciable extent, but that of wavemeter A was changed by about 2 parts in 10^4 when the length of the input cable was changed in steps by a total of one half-wavelength. Trouble was also experienced because of a resonance in the crystal detector circuit at a frequency near 9,500 Mc/s. This resulted in an effective increase in the coupling over this region of frequency and a consequent reaction on the resonant frequency of the wavemeter. The difficulty was overcome by making the insulating support of the crystal detector and coupling loop from a material of high dielectric loss.

To check the uniformity of the variation of the resonant frequency with movement of the plunger, the frequencies corresponding to a number of closely spaced settings were measured at one portion of the range of each instrument. The results obtained for wavemeter B are shown graphically in Fig. 4, and it is seen that no point lies off the straight line by more than 1 part in 10^4 of frequency. Similar results were obtained with the other wavemeters; but when one of the plungers was deliberately made eccentric so that the clearance gap between the plunger and wall varied with the rotation of the plunger there was a periodic variation of frequency of about $\pm$ 20 parts in 10^4 superimposed on the general line of the calibration curve.

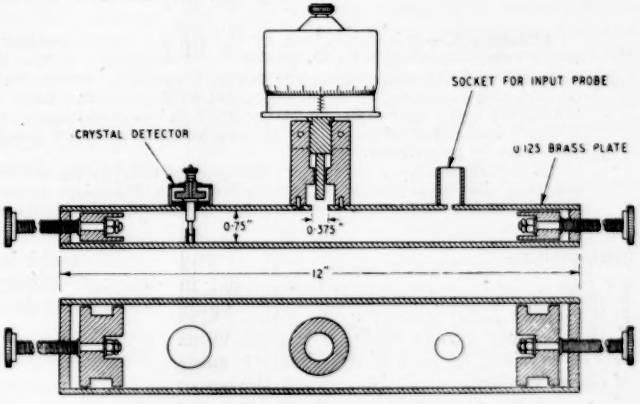


Fig. 6. *Wavemeter A is shown attached to the wall of a waveguide.*

In the first model of wavemeter B, the dimension l_2 was 2.54 cm, thus permitting the full movement of the micrometer. It was found, however, that a resonance effect occurred giving the irregularity in the calibration curve and decrease in the value of "Q" shown in Fig. 5. These effects were accompanied by a large decrease in sensitivity. When the dimension l_2 was decreased to 2.23 cm there was no such irregularity in the curve, but the values of Q were low at the highest frequencies, suggesting that such a resonance would have occurred at a frequency a little above the range of the wavemeter.

5. Modified Wavemeters B and C

There are two drawbacks to fixing an extension to the micrometer spindle as shown in Fig. 1. The first is that considerable care is required to make the extension and the spindle concentric, and the second is that the micrometer cannot be dismantled for cleaning. A micrometer was therefore made with the spindle extended by the appropriate amount, the extension being of slightly smaller diameter than the normal spindle. After being silver-plated the extension could then still be withdrawn from the barrel of the micrometer. Wavemeters were made with these micrometers, the gap between the spindle and the walls of the wavemeters again being 0.002 in. They gave a satisfactory performance with frequency ranges and Q-factors not very different from those previously obtained. They have some advantages, therefore, but this extension could not easily be fitted to a standard micrometer without softening the spindle.

6. The Use of Waveguide Feeds

In many applications at frequencies of 3,000 Mc/s, or over, the energy is transmitted through waveguides, and it is convenient to couple a wavemeter to the system by means of coupling holes in the guide and in the wall of the wavemeter. To check the use of the wide range wavemeters described, under such

conditions, the highest frequency model A was fitted on a guide as shown in Fig. 6, the dimensions of the guide being chosen to permit the propagation of TE_{11} waves at frequencies above 4,000 Mc/s. It was found that in order to obtain an absorption dip of convenient amplitude the diameter of the coupling window needed to be $\frac{3}{8}$ in.

The calibration curve of the wavemeter under these conditions was not very different from that shown in Fig. 3, except that the slope of the curve did not increase at the low frequency end. No troubles due to different input or detector conditions were experienced with this method of coupling.

7. Acknowledgements

The work described above was conducted as part of the programme of the Radio Research Board, to whom confidential reports were circulated during 1944. This paper is published by permission of the Department of Scientific and Industrial Research.

Some preliminary tests concerned with the development of the wavemeters were made by Dr. J. E. Johnston, and the calibration of the wavemeters was largely done by Mr. D. R. Dicks. The wavemeters were made in the workshop of the Radio Division of the National Physical Laboratory.

ABSOLUTE BELS*

By F. S. G. Scott

SUMMARY.—It is shown that, by the use of a simple operator and the choice of a suitable power reference level, not only power but many other factors (for example, dynes, lengths, resistances, volts, etc.) which are not of themselves power, but which condition power in certain circumstances, may be expressed in a particular form of power ratio described as "Absolute bels," that the term "Absolute bels" is justified, and that such an expression gives a direct indication of how these factors will condition the power in any circuit into which they may be introduced.

It is further shown that where the ratio of two powers is involved, this can be converted mentally and instantaneously into decibels or a difference in phons. Examples are given of the use of the absolute bel which indicate clearly the advantages which this unit possesses.

Introduction

IT has often been emphasized that the bel is a pure ratio between two powers and that if any power ratio is expressed in bels or decibels, no idea of the absolute value of either power is available, unless the value of one of the powers is stated, or by some implication is known.

To increase the value of this very useful mathematical device, it has been suggested that some standard power should be used for reference purposes, and certain values are, in fact, often used. Two examples are :—one milliwatt expended in 600 ohms, and six milliwatts expended in 500 ohms. These different standards tend, by their very variety, to defeat their own ends and cause confusion. Further, standards of this nature lead to endless difficulties if, for example, a decibel meter calibrated for one particular circuit, is required for use in another circuit with a different impedance.

* MS. accepted by the Editor, February 1946.

It should be, and indeed is, possible, to arrange matters in such a way that these difficulties do not occur.

Fundamental Considerations

If a power P_x is compared with a universally recognised and accepted standard power P and that answer is expressed in bels, then implicit in that answer is the absolute value of P_x and, in the same way

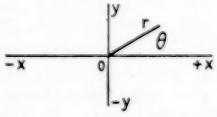


Fig. 1. *The line r is defined as* $r = |r|(\cos\theta + j\sin\theta)$.

that a line r having a defined position in space

$$r = |r|(\cos\theta + j\sin\theta) \quad \ldots \quad \text{(Fig. 1)}$$

gives more information than any simple line r and is known not as a line but as a vector, so the bel mentioned above gives more information than the power ratio and

is entitled to a distinguishing adjective to indicate this fact.

As "Absolute" values are determinable with such a convention, the adjective "Absolute" appears to be eminently suitable.

The above statement may be expressed mathematically as follows :—

If N_{p_x} be the value of P_x expressed in absolute bels then $\log_{10}(P_x/P) = N_{p_x}$ which may be rewritten

$$\log_{10} P_x - \log_{10} P = N_{p_x}$$

Now P is an arbitrary reference power, which should be standardized at the value which will offer maximum simplicity, and it is obvious that the complete elimination of the term $\log_{10} P$, by equating P to unity, satisfies this condition.

Definition

When any power (P_x) is compared (in bels) with one watt, the resultant answer is expressed in absolute bels. (B = 10 dB = *ten absolute decibels*).

$$N_{p_x} = \log_{10} P_x \quad .. \quad .. \quad .. \quad (1)$$

The practical meaning of this is best realized by an example. Assume that it is required, for correction purposes, to apply a "boost" at 50 c/s of 12 db to an amplifier, having a maximum rated output of 40 watts (this is described as 1.6 B in an absolute bel system) which is at present supplying 3 watts (0.48 B) at all frequencies. It is required to know :—

(*a*) Whether the amplifier has sufficient power output to meet this demand without overloading.

(*b*) What is the maximum permissible "boost."

The normal method of answering these questions is :—

(*a*) Let P_x be the power output at 50 c/s with 12 db "boost."

Then $10 \log_{10}(P_x/3) = 12$ db

$\therefore \quad \log_{10}(P_x/3) = 1.2 = \log_{10} P_x - \log_{10} 3$

or $\quad \log_{10} P_x = 1.2 + 0.48 = 1.68.$

Consequently,

$$P_x = \text{antilog}_{10}\, 1.68 = 48 \text{ watts.}$$

This is greater than 40 watts and the output of the amplifier is not sufficient to permit a "boost" of 12 db.

(*b*) Maximum permissible "boost" = $10 \log_{10}(40/3) = 10 \log_{10}(13.\dot{3}) = 10 \times 1.12 = 11.2$ db.

With the Absolute bel method the procedure is :—

(*a*) let N_{p_x} be the power output (in absolute bels) at 50 c/s with 12 db "boost." Then

$$N_{p_x} = 0.48 + 1.2 = 1.68 \text{ B}$$

The maximum rated output of 1.6 B is therefore inadequate.

(*b*) Maximum permissible "boost" = 1.6 B − 0.48 B = 1.12 b = 11.2 db.

Extension of the Principles of the Absolute Bel

In communications engineering it is often necessary to know what power will be developed under given conditions in a load which may be either acoustical or electrical.

It is, therefore, very convenient to express the value of any factor (F) or factors (F_1, F_2, F_3, . . . F_n) which will condition the power in a load, or to express the value of the load itself, in terms of the effect on the power developed in the load in absolute bels, and to describe such an expression as "the value of the factor or factors in Absolute Bels" $N_{(F)}$, or $N_{(F_1, F_2, F_3, \ldots F_n)}$.

The following definitions and conventions are used throughout this paper.

(1) $N_{(F)}$ = the value of any factor (F) expressed in absolute bels.

(2) Any impedance $R_s \pm jX_s$ (see Fig. 2). is described as $Z_s \underline{|\theta_s}$ the subscripts relating the various terms to each other.

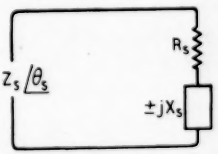


Fig. 2. *Any impedance $R_s \pm jX_s$ is expressed as $Z_s \underline{|\theta_s}$.*

The term $\underline{|\theta_s}$ is a simple method of remembering that strictly speaking

$$Z_s = Z_s(\cos\theta_s \pm j \sin\theta_s) = R_s \pm jX_s$$

Where $\theta = \tan^{-1}(X_s/R_s)$, and may be positive or negative. This ensures that junior engineers using these formulae are reminded that

Z_A added to Z_B is not $Z_A + Z_B$ but

$$(Z_A \cos\theta_A + Z_B \cos\theta_B) + j(Z_A \sin\theta_A + Z_B \sin\theta_B)$$

(3) Bels are written b. Decibels are written db.

Absolute Bels are written B. Absolute decibels are written dB.

(4) For brevity $\log_{10}$ is written A.

(5) The conversion of any factor (F) to the form $N_{(F)}$ is performed by means of an operator $O_{(F)}$.

Thus $N_{(F)} = O_{(F)}(F)$ .. .. .. (2)

We are now in a position to consider $O_{(F)}$ for various factors.

To determine O_{p_x} where P_x is any power (measured in watts).

From (2) $O_{p_x} P_x = N_{p_x}$

From (1) $N_{p_x} = AP_x$

$\therefore \quad O_{p_x} P_x = AP_x$

$\therefore \quad O_{p_x} = A = \log_{10}$ .. .. (3)

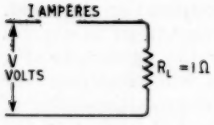


Fig. 3. *The power in a load resistor is expressed relatively to a standard power of* 1 *watt.*

In Fig. 3 the power P_L developed in the load resistor R_L of 1Ω may be expressed

$$P_L = V^2/1$$

Now by the definition given for $N_{(F)}$ and from equation (2)

$$AP_L = N_V = O_V V$$

But $P_L = V^2$

$\therefore \quad AP_L = AV^2 = 2AV = O_V V$

and $O_V = 2A = 2 \log_{10}$

Thus $N_V = 2AV$ .. .. .. (4)

Similarly

$$P_L = I^2 \times 1$$

and $AP_L = N_I = O_I I$

$\therefore \quad AP_L = 2AI = O_I I$

$\therefore \quad O_I = 2A = 2 \log_{10}$

Thus $N_I = 2AI$ .. .. .. (5)

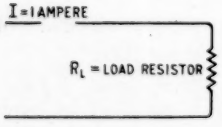


Fig. 4. *Development of the load factor* N_{R_L} *in terms of current.*

In Fig. 4

$$P_L = 1^2 \times R_L = R_L$$

Now by definition

$$N_{R_L} = O_{R_L} R_L = AP_L$$

But $P_L = R_L$

$\therefore \quad O_{R_L} = A = \log_{10}$

Thus $N_{R_L} = AR_L$ .. .. .. (6)

In Fig. 5

$$I = 1/Z_L$$

$\therefore \quad P_L = 1^2/(Z_L)^2 \times 1 = 1/Z_L^2 = Z_L^{-2}$

By definition

$$N_{Z_L} = O_{Z_L} Z_L = AP_L$$

But $P_L = Z_L^{-2}$

$\therefore \quad O_{Z_L} Z_L = -2AZ_L$

$\therefore \quad O_{Z_L} = -2A = -2 \log_{10}$

Thus $N_{Z_L} = -2AZ_L$ .. .. .. (7)

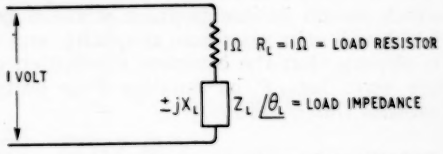


Fig. 5. *Illustrating a load circuit containing both reactive and resistive elements.*

To indicate clearly fundamental principles, up to this point only individual factors have been considered, but this technique is not necessary.

Let us now deal with a case of some importance illustrated in Fig. 6.

V_G is the "open circuit" voltage of the generator. By definition

$$N_{(V_G Z_G Z_L R_L)} = O_{(V_G Z_G Z_L R_L)} f(V_G Z_G Z_L R_L) = AP_L$$
.. .. (8)

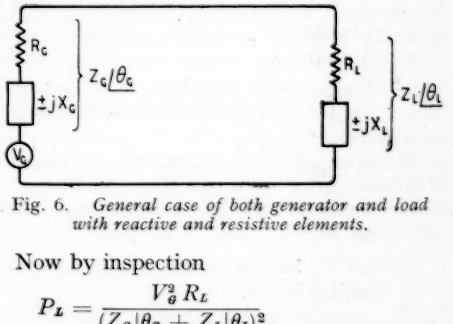


Fig. 6. *General case of both generator and load with reactive and resistive elements.*

Now by inspection

$$P_L = \frac{V_G^2 R_L}{(Z_G \underline{|\theta_G} + Z_L \underline{|\theta_L})^2}$$

$$\therefore AP_L = A \left[\frac{V_G^2 R_L}{(Z_G \underline{|\theta_G} + Z_L \underline{|\theta_L})^2} \right]$$

i.e. $N_{(V_G Z_G Z_L R_L)} = 2AV_G + AR_L - 2A \left[(Z_G \underline{|\theta_G} + Z_L \underline{|\theta_L}) \right]$ .. (9)

Considering the first two terms on the right-hand side of the equation we notice that, as we should expect, volts and the load resistance

are expressed in absolute bels in the form deduced in equations (4) and (6) and, further it is obvious that

$$O_{(F_1,F_2,F_3\ldots F_n)} = +O_{(F_1)}+O_{(F_2)}+O_{(F_3)}\ldots+O_{(F_n)}$$

The third term on the right of equation (9) is known as the matching factor (M) for it indicates the effect that the relative values of the load impedance and the generator impedance have upon the final power produced.

A useful " particular " case to remember is that where the load is " matched " to the generator, i.e. $R_L = R_G$ and $\pm jX_L = \mp jX_G$ then (since we have optimum matching N_M is written N_{OM})

$$N_{OM} = -2A[2R_L] = -2[AR_L + 0.3] \quad .. \quad .. \quad (10)$$

a further useful case to consider, is that in which the voltage across the load impedance is known ; this is really the case of

$$Z_G = O\angle\theta^\circ$$

and equation (9) becomes

$$AP_L = 2AV_G + AR_L - 2AZ_L\angle\underline{\theta_L} \quad .. \quad .. \quad (11)$$

which, if the load is purely resistive, simplifies to

$$AP_L = 2AV_G + AR_L - 2AR_L = 2AV_G - AR_L \quad .. \quad .. \quad (12)$$

General Applications of B

It should now be clear that a long " chain " network may be much simplified by the following technique. Let Fig. 7 represent any such network.

It is obvious that if all terms are expressed in absolute bels

$$N_{F_L} = 2AV + AR_L + N_{F_1} + N_{F_2} + N_{F_3} \ldots + N_{F_n} \quad .. \quad (13)$$

Thus all the essential facts regarding an amplifier (impedances, rated output, etc.) should be expressed in absolute bels. It may be noticed in passing that while this arrangement presents all data in their most simple form for speedy calculations, information in more normal notation is readily obtainable since to determine (F) from $N_{(F)}$ only requires one simple and obvious operation.

In Fig. 8 a " transformer " is inserted between a generator $Z_G\angle\underline{\theta_G}$ and a load $Z_L\angle\underline{\theta_L}$ which, with both generator and load connected, " looks like " $Z_P\angle\underline{\theta_P}$ to the generator and $Z_S\angle\underline{\theta_S}$ to the load. This " transformer " may be a transformer, as the term is normally understood, in which case it will transmit x per cent. of the power applied to it, or the " transformer " may be an amplifier, in which case (neglecting the " raw " power supplied to it) it will transmit more power than it receives.

These transformers have a gain factor G which in the case of an amplifier having a gain control, is marked upon the control in such a manner that, in the position of maximum gain, the value of N_G is shown, while in all other positions, the effect of the gain control upon the N_G figure is taken into account.

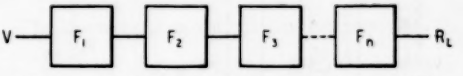


Fig. 7. *A chain network is represented symbolically here.*

Thus in an amplifier having a gain control in 3 db stages, with a maximum gain in the, clockwise direction, and an N_G of 10 dB, the gain control would be marked in the way indicated in Fig. 9.

N_G is determined in the following manner :

Let $\dfrac{mVZ_P\angle\underline{\theta_P}}{Z_G\angle\underline{\theta_G} + Z_P\angle\underline{\theta_P}}$ be the " generator " volts in the secondary circuit (see Fig. 8) when

$$\frac{VZ_P\angle\underline{\theta_P}}{Z_G\angle\underline{\theta_G} + Z_P\angle\underline{\theta_P}} \text{ Volts}$$

are applied across the primary $Z_P\angle\underline{\theta_P}$

Then

$$P_L = \frac{\left[\dfrac{mVZ_P\angle\underline{\theta_P}}{Z_G\angle\underline{\theta_G} + Z_P\angle\underline{\theta_P}}\right]^2}{\left(Z_S\angle\underline{\theta_S} + Z_L\angle\underline{\theta_L}\right)^2} R_L$$

$$\therefore \quad AP_L = 2AV + AR_L - 2A\left(Z_G\angle\underline{\theta_G} + Z_P\angle\underline{\theta_P}\right) - 2A\left(Z_S\angle\underline{\theta_S} + Z_L\angle\underline{\theta_L}\right) + 2A(Z_P m) \quad .. \quad .. \quad (14)$$

$2A(Z_P m)$ is the gain factor (N_G)

and we see that equation (14) is in what is now rapidly becoming a familiar form (reading the right side from left to right) with an E.M.F. factor, load factor, primary and secondary matching factors and a gain factor to indicate what is occurring in the " transformer."

If $Z_G\angle\underline{\theta_G} = Z_P\angle\underline{\theta_P} = Z_S\angle\underline{\theta_S} = Z_L\angle\underline{\theta_L} = R_L$

equation (14) reduces to

$$AP_L = 2AV - 3AR_L + N_G - 1.2 \quad .. \quad .. \quad (15)$$

A loudspeaker is, of course, a " particular " case of the transformer problem, in this case transforming electrical into acoustical power. Consider equation (14). The terms $-2A(Z_S|\underline{\theta_S} + Z_L|\underline{\theta_L})$ and $2A(Z_P m)$ are electro-acoustical and obviously depend upon the construction of the instrument, while AR_L (a purely acoustical term) is for practical purposes a constant. (See note on K below). They may therefore be " lumped " and described as the transformation factor of the loudspeaker (T).

does not from a practical viewpoint convey the desired information.

What is of interest is not what is happening " in the box " but what is happening at a point a reasonable distance from it.

With the types of loudspeaker normally used the sound is diverging at the distance from the instrument at which one may expect to find a listener and, generally it is found, in the writer's experience, that anyone who is attempting to obtain optimum " listening " tends to sit on, or within a

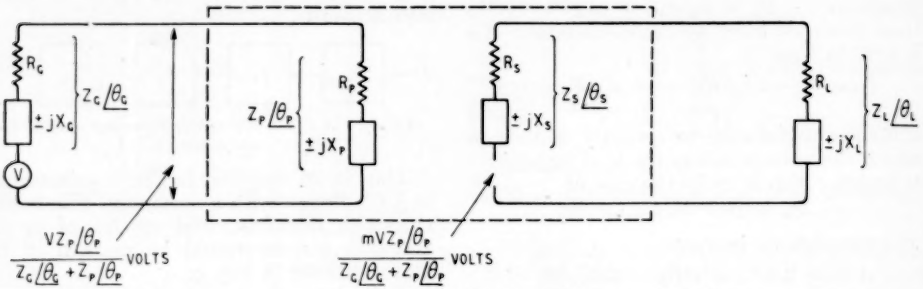


Fig. 8. *A transformer or amplifier represented by Z_p and Z_s is connected between generator and load.*

Then if P_A is the total acoustical power produced by the instrument

$$AP_A = 2AV - 2A(Z_G|\underline{\theta_G} + Z_{LS}|\underline{\theta_{LS}}) + N_{T_{LS}} \quad .. \quad .. \quad (16)$$

Where $Z_{LS}|\underline{\theta_{LS}}$ is the electrical input impedance of the unit. Let us suppose that we are able to determine experimentally that when R_{LS} is the " in phase " component of a loudspeaker the application of an electrical power

$$P_E = \frac{V^2 R_{LS}}{(Z_G|\underline{\theta_G} + Z_{LS}|\underline{\theta_{LS}})^2}$$

produces a total acoustical power P_A and let $P_E = \beta P_A$.

Then

$$AP_A = 2AV + AR_{LS} - 2A(Z_G|\underline{\theta_G} + Z_{LS}|\underline{\theta_{LS}}) - A\beta \quad (17)$$

From 16 and 17.

$$2AV - 2A(Z_G|\underline{\theta_G} + Z_{LS}|\underline{\theta_{LS}}) + N_{T_{LS}} = 2AV + AR_{LS} - 2A(Z_G|\underline{\theta_G} + Z_{LS}|\underline{\theta_{LS}}) - A\beta$$

$$\therefore \quad N_{T_{LS}} = AR_{LS} - A\beta \quad .. \quad .. \quad (18)$$

To determine P_A (and hence β the " badness " of the loudspeaker) accurately presents some difficulties and the resultant β reasonable angle off, the centre line which is at right angles to the " plane of diffusion " of the machine.

In these conditions no appreciable error is involved in assuming that the power is varying inversely with the square of the distance from the " front " of the loudspeaker.

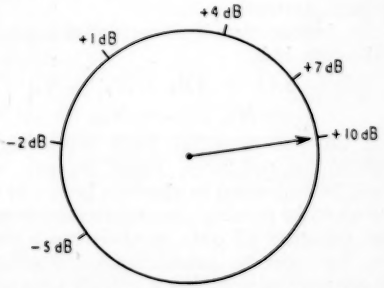


Fig. 9. *Scale of gain control calibrated in absolute decibels.*

We can then make an arbitary and untrue assumption (which does not introduce any error in determining the acoustical pressure at a listener's ear) that the loudspeaker is radiating all the power evenly through a solid angle of $2\cos^{-1}\frac{2\pi - 1}{2\pi}$ (about 66°)

so that the acoustical power per unit area at any distance l is inversely proportional to l^2.

Then if the acoustical power per unit area (P_{AT}) at a suitable test distance l_t be measured by a convenient means when a known electric power P_E is applied to the instrument

$$A\beta = AP_E - AP_A = AP_E - AP_{AT} - 2Al_t \quad .. \quad .. \quad (19)$$

We see, therefore, that for acoustical purposes any length l may be expressed in absolute bels (N_l) from the formula

$$N_l = 2Al$$

This convention does not of course give any indication of the actual total acoustical power produced by the instrument, but it does give an accurate indication of the usable power, which is what is of interest.

The above principles may with advantage be applied in reverse to microphone problems.

The Phon and Dyne/cm²

The phon, although expressed in decibel notation, is of course more than a power ratio as all powers are referred to a known power of 10^{-16} watts.

In other words, it would be an " Absolute bel expression " if only the reference level were different (10° watts) and phons at 1,000 c/s may be converted to dN_{P_h} by the simple deduction of the constant 160.

Thus at 1,000 c/s x/10 phons $-$ 16 = Acoustic power expressed in absolute bels.

At any frequency other than 1,000 c/s it is necessary to add a correction C (in bels) to provide for the characteristics of the human ear as indicated in the curves by Fletcher and Munson.

If P_A = acoustical power expressed in watts/cm²

d = dynes/cm²

K = a constant for any given conditions

Then $$\frac{d^2}{K} = P_A$$

$$\therefore \quad AP_A = 2Ad - AK \text{ or } AK = 2Ad - AP_A \quad .. \quad .. \quad (20)$$

Now under normal conditions of loudspeaker working a force of 204×10^{-6} dynes/cm² produces a power of 10^{-16} watts.

From equation (20)

$$AK = 2(\overline{4}.31) + 16 = -7.38 + 16 = 8.62$$

A summary of all the results appears in Table 1.

As an example, it is required to know what voltage and what electrical power should be applied to a speech coil (having a resistance of 0.3 B and impedance of $-$ 0.8 B both at 50 c/s, a transformation factor $-$ 2.2 B and a frequency discrimination compared with 1,000 c/s of $-$ 1 b at 50 c/s), in order to produce 75 phons at 50 c/s at a distance of 184 cm, from the cone on the centre line.

A study of the Fletcher and Munson curves indicates that it requires 0.7 bels more power to obtain 75 phons at 50 c/s compared with the power required at 1,000 c/s, i.e. $C = 0.7$ b.

The acoustical power at the loudspeaker is

$$\frac{75}{10} - 16 + 0.7 + 4.53 = AP_A = -3.27 \text{ B}$$

The first three terms are concerned with the conversion from phons to watts, and the fourth with the distance of 184 cm.

Now $N_{P_A} = N_V + N_{Z_{LS}} + N_T - 1$

(Compare with equation 16).

$$\therefore \quad N_V = 1 - N_{Z_{LS}} - N_T + N_{P_A}$$
$$= 1 + 0.8 + 2.2 - 3.27 = 0.73 \text{ B}$$

$$V = \text{Antilog}_{10} \frac{0.73}{2} = 2.3 \text{ volts.}$$

and

$$N_{P_E} = N_V + N_{R_{LS}} + N_{Z_{LS}} = 0.73 + 0.3 - 0.8 = 0.23 \text{ B}$$

$$\therefore \quad P_E = 1.7 \text{ watts.}$$

Where it is desired to ascertain relative as well as quantitative readings on linear volt or ampere meters, they may be marked as in Fig. 10. This scale is one which the author finds most useful for his particular

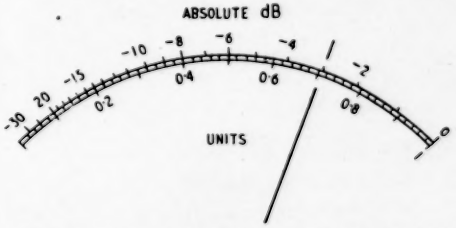


Fig. 10. *Metre scale with calibration in absolute decibels.*

class of work. He is always interested in two things, viz. :

(1) what is happening in the circuit to which the meter is attached and,

(2) what is happening (in phons) in space 200 cm, from standard types of loudspeakers

which may be located anywhere in a very large area.

Table 2 indicates the relation between the dB and Unit scales. The long mark right through the dial (Fig. 10) indicates the normal maximum reading for the instrument. This gives a margin on the meter of 3 db in hand, so that, if at any time, through bad control, the power is being transmitted at too high a level, there is a direct indication of the value of this overload and also a margin of safety as far as the meter itself is concerned.

It will be realised that the scale is correct only for a maximum R.M.S. voltage or current of unity applied to a non-reactive resistance of 1 ohm, but it holds good wherever it may be installed, if read in conjunction with a small correction card, which is attached to the outside of the meter and, changed whenever the meter is installed in a different circuit.

Fig. 11 shows a typical card, and this paper may well close by showing the extreme ease and speed with which the data for this card may be obtained. The case considered is that of the network shown in Fig. 12. The card is for use with a voltmeter which is across the input terminals of an amplifier (No. 1) which normally

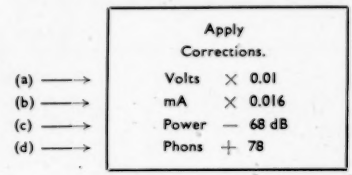


Fig. 11. *Typical correction card for a meter scale.*

TABLE I

POWER (P) $N_p = AP$

Volts (V)	$N_V = 2AV$	**Current (I)**	$N_I = 2AI$
Resistance (R)	$N_R = AR$	**Impedance (Z)**	$N_Z = -2AZ$
Matching Factor (M) (Where $Z_G \lvert\underline{\theta_G}$ and $Z_L \lvert\underline{\theta_L}$ are generator and load impedances, or sending and receiving impedances respectively).	$N_M = -2A$ $(Z_G \lvert\underline{\theta_G} + Z_L \lvert\underline{\theta_L})$	**Optimum Matching** $(R_G \pm jX_G = R_L \pm jX_L)$	$N_{OM} = -2[AR_L + 0.3]$
Gain Factor (G) (Where Z_p is the primary impedance and m equals the volts appearing across the secondary when it is "on open circuit" divided by the volts at that time being applied across Z_p).	$N_G = 2A\,[Z_p m]$	**Distance (l)**	$N_l = 2Al$
		Dynes/cm (d)	$N_d = 2Ad - 8.62$
Loudspeaker transformation factor (T) (Where R_{LS} equals the input resistance of the loud speaker).	$N_T = AR_{LS} - A\beta$	**Effective badness** (β) of Loudspeaker. (Where P_E equals electric power applied to the loudspeaker for an accoustic power P_{AT} per cm² at a test distance l_T.)	$N_\beta = AP_E - AP_{AT} - 2Al_T$

CONVERSIONS

Dynes/cm² (d) to phons (Ph)	$2Ad + 7.38 - C = \frac{\text{Ph}}{10}$	**Phons (Ph) to dynes/cm² (d)**	$\frac{\text{Ph}}{10} + C - 7.38 = 2Ad$
Phons (Ph) to watts/cm² (P_A)	$\frac{\text{Ph}}{10} + C - 16 = AP_A$	**Watts/cm² (P_A) to dynes/cm² (d)** (Where C is the additional power (in bels) required at any frequency in accordance with the Fletcher and Munson Curves).	$AP_A + 8.62 = 2Ad$
Watts/cm² (P_A) to phons (Ph)	$AP_A + 16 - C = \frac{\text{Ph}}{10}$		

receives a maximum R.M.S. voltage of 0.007 volts. This amplifier feeds an identical amplifier (No. 2) via a line having an insertion loss of 25 db.

It is a convenient and simple practice with absolute bels to assume that any two pieces of equipment actually joined by a line are directly joined and to describe the line as having an insertion loss in accordance with the facts.

Amplifiers (1) and (2) are indentical line amplifiers, each having resistive inputs and outputs of 2.8 B and an N_G of 10.5 B, the gain control in amplifier (1) being in the full gain position, and that in amplifier (2) is applying an insertion loss of 18 db.

TABLE II

dB	Units	dB	Units
1	0.89	12	0.25
2	0.79	13	0.22
3	0.71	14	0.20
4	0.63	15	0.18
5	0.56	16	0.16
6	0.50	17	0.14
7	0.45	18	0.12
8	0.40	19	0.11
9	0.36	20	0.10
10	0.32	25	0.06
11	0.28	30	0.03

Amplifier (2) feeds a third amplifier (3) having a resistive input of 2.8 B, an N_G of 9.3 B and an output impedance of $3 + j0$ ohms (0.48 B). Its gain control is set to an insertion loss of 5 db. This amplifier in turn feeds a loudspeaker having an input impedance of $8000 + j6{,}000$ (− 8 B) a transformation factor (N_T) of 1.0 B fed by a line having a 1-db insertion loss. The loudness in phons is to be computed at a distance of 200 cm on the centre line from the instrument.

Since normal maximum volts = 0.007 (R.M.S.) (read at normal maximum line 3 db below full scale.)

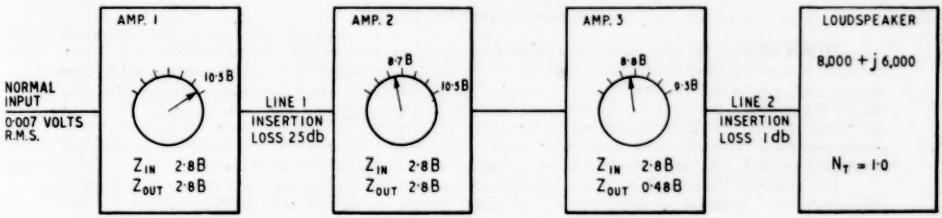


Fig. 12. *Amplifier arrangement used to illustrate the application of absolute bels.*

Full-scale reading $N_F = 0.3 + 2A\ 0.007 = 0.3 - 4.3 = -4$B, i.e. 0.01 volt R.M.S. .. .. (a)

From first principles

$2AI = -4 - 5.6 = -9.6$ B

$\therefore \quad AI = -4.8$B, i.e. 0.016 mA .. (b)

and $N_P = -4 - 2.8 = -6.8$B .. (c)

$\frac{\text{Phons}}{10}$ equals

$+ 16$ = w/Ph conversion constant
$- 4$ = Applied volts
$- 5.6$ = Input match amp. 1
$+ 10.5$ = N_G amp. 1
$- 2.5$ = Insertion loss line 1
$- 6.2$ = Matching amp. 1 to amp. 2
$+ 8.7$ = N_G amp. 2 corrected for attentuator setting
$- 6.2$ = Matching amp. 2 to amp. 3
$+ 8.8$ = N_G amp. 3, corrected for attenuator setting
$- 0.1$ = Insertion loss line 2
$- 8$ = Matching amp. 3 to loudspeaker
$+ 1.0$ = N_T
$- 4.6$ = 200 cm

$= 7.8$

i.e. 78 phons .. .. (d)

BEAM TETRODE CHARACTERISTICS*

The Effect of Electron Deflections

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SUMMARY.—In this paper the anode-current—anode-voltage characteristic of a beam tetrode is deduced on the assumption that the electrons entering the screen-to-anode space are not all projected normally to the screen plane, but that they possess a distribution-in-angle given by a particular continuous function, for convenience given in terms of the sine of angle of entry. This supplements previous treatments of the problem, in which it has been assumed that above a maximum angle the relative number of entering electrons falls abruptly to zero, i.e. the distribution is represented by a discontinuous function.

It is shown that, although this discontinuous feature is lacking in the newly assumed distribution, "regions of instability" yielding sharp knees are still obtained, due to the action of space charge.

Introduction

IT has been recognised for some time that the anode-current—anode-voltage characteristics of beam tetrodes are greatly modified by the fact that the electrons which enter the screen-to-anode space have previously been deflected at the grid and screen wires [1–5]. The main consequence of such deflections is that the forward electron velocities are diminished, so that the strongly deflected electrons are brought to rest and then reflected back to the screen from some plane intermediate between the screen and anode where the potential is sufficiently low, but still positive. If the current were carried by undeflected electrons such reflection could occur only at a "virtual cathode," where the potential is zero, but the "virtual cathode" theory leads to tetrode knee-voltages which are much too high.

In this paper attention is restricted to the plane parallel electrode arrangement, with an infinitely long and infinitely wide emitting cathode, so that if a section of the beam is taken at a plane parallel to the cathode, the area of the beam is infinitely large. In this case edge effects can be neglected. It is assumed that the electron trajectories at entry into the screen-to-anode space are inclined to the normal to the screen plane, and it is further necessary to make some assumption regarding the distribution-in-angle of the current projected into this space. For the calculations which follow a modified expression for the distribution function has been adopted; this gives the current density distribution as a function of the sine s of the angle of entry; thus

$$dI = I_T i(s)\, ds$$

where $I_T i(s)$ is the current density comprised between values of the sine of the angle of entry ranging from s to $s + ds$.

I_T is the total current density, i.e. the current flowing across unit area of the screen plane.

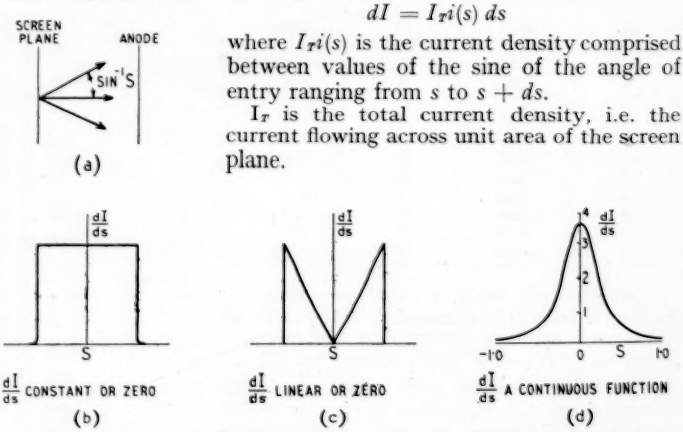

Fig. 1. *Various assumed distributions plotted against sine of angle of entry.*

dI/ds [i.e. $I_T i(s)$] is the distribution function as shown in Fig. 1. When $i(s)$ is known, the potential throughout the screen-to-anode space can be precisely evaluated. This has already been done, for several simple assump-

* MS. accepted by the Editor, November 1945.

tions regarding $i(s)$, by Strutt and van der Ziel, by G. B. Walker and by the writer. It is the object of the present paper to extend these calculations to the important case where $i(s)$ is a continuous, instead of a discontinuous function, and to show that "regions of instability" [4] yielding sharp knees, are still obtained.

The Current reaching the Anode

Let it be supposed that the screen plane is at a potential V_1 above the cathode, with the anode at a potential V_2, and that there is a potential minimum, of magnitude V_m, somewhere between the screen and anode.

[In what follows we shall write

$$V/V_1 = W\ ;\ \sqrt{V/V_1} = w\ ;\ \sqrt{V_m/V_1} = w_m\ ;$$
$$\sqrt{V_2/V_1} = w_2\ ;\ \text{and}\ \sqrt{V} = v]$$

The condition that an electron should reach the anode is that $V_1 s^2$ should be less than V_m, i.e. s lies between $-w_m$ and $+w_m$. The primary current incident on the anode is therefore

$$I_A = I_T \int_{-w_m}^{+w_m} i(s)\, ds \quad .. \quad .. \quad .. \quad (2)$$

This equation for I_A gives the anode current as a function of w_m; for fixed conditions of grid and screen voltage this characteristic should be independent of the screen-to-anode gap. For small values of I_T, or for short gaps, a potential minimum will not exist, but if the anode voltage is lower than the screen voltage the anode voltage V_2 will be the lowest in the screen-to-anode space, and hence w_2 should replace w_m as the limits in the integral. With regard to the form of $i(s)$ it may be shown that when the deflections at the grid and screen wires are cumulative, as in a tetrode, the basic I_A—w_m characteristic approaches the full value gradually, and therefore the function $i(s)$ should represent a continuous curve [5].

In order to reproduce this the following function has been chosen for $i(s)$

$$i(s) = \frac{1}{2\tan^{-1} k} \cdot \frac{k}{1 + k^2 s^2} \quad .. \quad (3)$$

where k is an arbitrary constant.

This yields on integration with respect to s, with limits $-w_m$, $+w_m$

$$I_A = I_T \left[\frac{\tan^{-1} k w_m}{\tan^{-1} k} \right] \quad .. \quad .. \quad (4)$$

These relations are shown in Figs. 1(*d*) and 2.

The Space Charge Conditions between Screen and Anode

If the space-charge density is known throughout the screen-to-anode space we can find the potential distribution by effecting a solution of Poisson's equation.

Now if dI is the current density comprised between s and $s + ds$ the corresponding contribution $d\rho$ to the negative space-charge density ρ is dI/u, where u is the forward electron velocity.

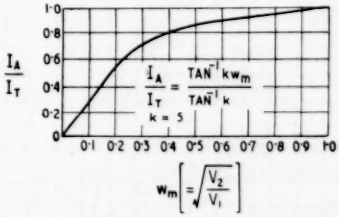

Fig. 2. *I_A/I_T plotted against w_m for assumed distribution.*

At a plane where the potential is V (E.S.U.) the value of u is

$$\sqrt{2\epsilon/m}\,\sqrt{V - V_1 s^2}$$

where ϵ = charge of electron in E.S.U.

m = mass of electron

But $dI = I_T\, i(s)\, ds$

$$\therefore \quad d\rho = \frac{I_T\, i(s)\, ds}{\sqrt{2\epsilon/m}\,\sqrt{w^2 - s^2}} \quad .. \quad .. \quad (5)$$

If sufficient current density is projected across the screen plane a potential minimum will be formed. On the anode side of the potential-minimum plane there are present only electrons for which $V_1 s^2$ is less than V_m, but on the screen side of the potential-minimum plane the circumstances are more complicated, since there are present both forward and returning electrons. The forward electrons, crossing a plane where the potential is V, comprise those for which $V_1 s^2 < V$, while the returning electrons comprise those for which $V_1 s^2$ lies between V_m and V (the electrons for which $V_1 s^2 < V_m$ go on to the anode).

Hence the value of ρ at a plane lying on the screen side of the potential minimum is given by

$$\sqrt{2\epsilon/m\, V_1}\,.\,\rho = I_T \left[\int_{-w}^{+w} \frac{i(s)\, ds}{\sqrt{w^2 - s^2}} + \int_{-w}^{-w_m} \frac{i(s)\, ds}{\sqrt{w^2 - s^2}} + \int_{w_m}^{w} \frac{i(s)\, ds}{\sqrt{w^2 - s^2}} \right] \quad (6)$$

If $i(s)$ is an even function, which is the practical case, this simplifies to

$$\sqrt{2\epsilon/m\,V_1}\,.\,\rho = 2I_T\left[2\int_0^w \frac{i(s)\,ds}{\sqrt{w^2-s^2}} - \int_0^{w_m}\frac{i(s)\,ds}{\sqrt{w^2-s^2}}\right] \quad .. \quad .. \quad (7)$$

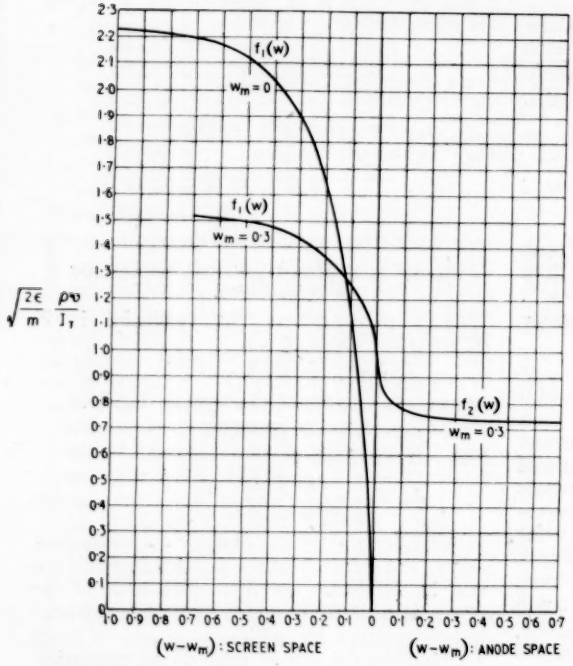

Fig. 3. $\sqrt{2\epsilon/m}\ \rho v/I_T$ *plotted against* $(w - w_m)$.

In a similar manner the space charge density at a plane on the anode side of the potential minimum is given by the expression

$$\sqrt{2/\epsilon m\,V_1}\,.\,\rho = 2I_T\int_0^{w_m}\frac{i(s)\,ds}{\sqrt{w^2-s^2}} \qquad (8)$$

It is interesting to consider the product ρv instead of the space charge density ρ; thus if each side of the above expressions is multiplied by w, we have, since

$$\sqrt{V_1}\,.\,w = v$$

$$\sqrt{2\epsilon/m}\,.\,\rho v = 2I_T\left[2w\int_0^w \frac{i(s)\,ds}{\sqrt{w^2-s^2}} - w\int_0^{w_m}\frac{i(s)\,ds}{\sqrt{w^2-s^2}}\right] \quad .. \quad .. \quad .. \quad .. \quad (9)$$

and $$\sqrt{2\epsilon/m}\,.\,\rho v = 2I_T\left[w\int_0^{w_m}\frac{i(s)\,ds}{\sqrt{w^2-s^2}}\right] \quad .. \quad .. \quad (10)$$

on the screen and anode sides of the potential minimum, respectively.

When $i(s) = \frac{1}{2\tan^{-1}k}\cdot\frac{k}{1+k^2s^2}$ the integrals can be evaluated (see Appendix) and give the following results :—

(1) *On screen side of potential minimum*

$$\sqrt{2\epsilon/m}\,.\,\rho v = I_T\frac{\sin\alpha}{\tan^{-1}k}[\pi - (\psi+\chi)] = I_T f_1(w), \text{ say} \quad .. \quad .. \quad (11)$$

(2) *On anode side of potential minimum*

$$\sqrt{2\epsilon/m}\,.\,\rho v = I_T\frac{\sin\alpha}{\tan^{-1}k}[\psi+\chi] = I_T f_2(w), \text{ say} \quad .. \quad .. \quad (12)$$

where

$$\alpha = \tan^{-1}kw\,; \quad \psi = \sin^{-1}w_m/w\,;$$

$$\chi = \tan^{-1}\left[\frac{\sin 2\psi\,\tan^2(\alpha/2)}{1-\cos 2\psi\,\tan^2(\alpha/2)}\right]$$

The courses of $f_1(w)$ and of $f_2(w)$ are shown in Fig. 3, with k chosen to be 5.

Determination of the Potential Distribution, knowing ρv

Poisson's Equation is $d^2V/dx^2 = 4\pi\rho$ where ρ is the density of negative space charge in E.S.U., and x is measured in cm. If both sides are multiplied by $2dV/dx$ (i.e. by $4v\ dv/dx$) we get

$$dV/dx\ (dV/dx)^2 = 16\pi\,\rho v\,dv/dx = 16\pi\,V_1^{\frac{1}{2}}\rho v\,dw/dx$$

$$\therefore \quad (dV/dx)^2 = 16\pi\,V_1^{\frac{1}{2}}\int_{w_m}^{w}\rho v\,.\,dw$$

or since

$$W = V/V_1$$

$$(dW/dx)^2 = \frac{16\pi}{V_1^{3/2}} \int_{w_m}^{w} \rho v \, . \, dw$$

Now ρv is of the form $\frac{I_T}{\sqrt{2\epsilon/m}} f(w)$

$$\therefore \quad (dW/dx)^2 = \frac{16\pi \, I_T}{\sqrt{2\epsilon/m} \; V_1^{3/2}} \int_{w_m}^{w} f(w) \, dw$$

In order to simplify this, we may note that the current which flows in a diode of gap x_a with anode maintained at a potential V_1 is

$$\frac{\sqrt{2\epsilon/m} \; V_1^{3/2}}{9\pi \; x_a^2} \text{ E.S.U/cm}^2 = I_d, \text{ say,}$$

If then I_T is put equal to $J_T \, . \, I_d$

$$(dW/dx)^2 = 16/9 \; J_T/x_a^2 \int_{w_m}^{w} f(w) \, dw$$

and

$$\therefore \quad \frac{x - x_m}{x_a} = \frac{3}{2\sqrt{J_T}} \int_{w_m}^{w} \frac{w \, dw}{\sqrt{\int_{w_m}^{w} f(w) \, dw}} \qquad .. \quad .. \quad (13)$$

From this the potential distribution against x can immediately be derived. Thus if the distance from the potential minimum to the screen is x_1

$$x_1/x_a = \frac{3}{2\sqrt{J_T}} \int_{w_m}^{1} \frac{w \, dw}{\sqrt{\int_{w_m}^{w} f(w) \, dw}} \qquad .. \quad (14)$$

the upper limit being unity, since $w = 1$ at the screen.

Again, when the total gap x_a from screen to anode is fixed, the potential minimum to anode distance x_2 must equal $(x_a - x_1)$ The value of the upper limit w must then be found which gives this value for x_2; thus the anode voltage V_2 is such that

$$x_2/x_a = \frac{x_a - x_1}{x_a} = \frac{3}{2\sqrt{J_T}} \int_{w_m}^{w_2} \frac{w \, dw}{\sqrt{\int_{w_m}^{w} f_2(w) \, dw}} \qquad .. \quad .. \quad (15)$$

The integrals must be evaluated by numerical methods; when this is done, there is at our disposal complete means for finding the anode-current—anode-voltage characteristics.

In order to do this, we first plot w_m against $\sqrt{J_T} \; x_1/x_a$ the values being computed from Equ. (14), This gives the boundary curve shown in Fig. 4, and indicates the relative position of the potential minimum plane. The course of w for any assigned w_m is next calculated from Equ. (13), with $f(w) = f_2(w)$, and plotted on the same graph against $\sqrt{J_T} \; x/x_a$. This is done for a suitable set of values of w_m. Now draw a vertical line displaced $\sqrt{J_T}$ from the origin; this will cut the family of w curves, and will give the values of w_2 corresponding to the values of w_m. Since, however, each w_m corresponds to a specific value of J_A/J_T we can plot J_A/J_T as a function of w_2 (Fig. 5), or better still, plot J_A against V_2/V_1. Fig. 6 shows J_A against V_2/V_1 for $J_T = 2$ and $J_T = 4$, and these are the computed shapes of the anode-current — anode-voltage characteristics.

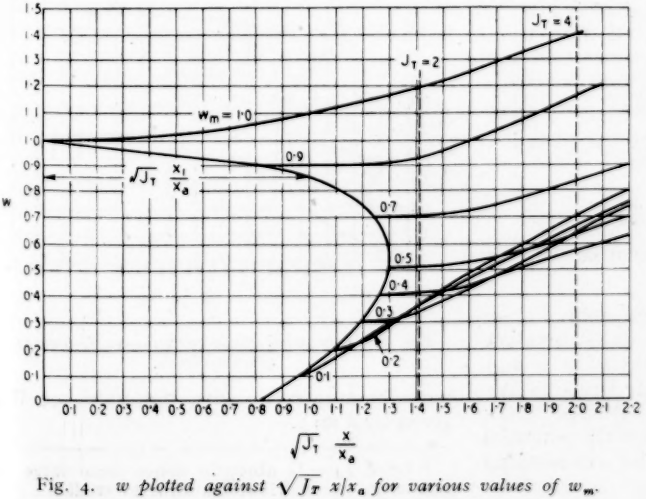

Fig. 4. *w plotted against* $\sqrt{J_T} \; x/x_a$ *for various values of* w_m.

In some ways the computed curves differ notably from the measured characteristics. It would be found in practice that the initial rising part of the $J_T = 4$ curve would approximately coincide with the initial rising part of the $J_T = 2$ curve. The writer does not think that this difficulty can

be overcome as long as attention is confined to beams which are of large cross-sectional area in relation to the length. The negative resistance region, from W_σ to W_K, also is not so well marked in practice, although the space charge due to secondary

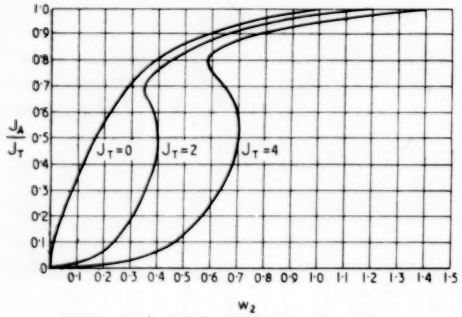

Fig. 5. *J_A/J_T plotted against w_2.*

electrons from the anode helps to reduce the effect.

If we identify W_σ the beginning of "the region of instability," as the beam tetrode "knee voltage" (since here, with increasing anode voltage, the anode current will shoot sharply up to a greater value), the calculations once more strikingly confirm that there is a large reduction in the knee voltage if the electrons are projected into the screen to anode space over a range of angles, as compared with the case when they all projected normally through the screen plane. The curves of Fig. 6 also verify the thesis of this paper that "knees" are obtainable even if the $i(s)$ characteristic is itself free from kinks.

Simplification where k is large

When k is sufficiently large, the equation for χ reduces to

$$\chi = \tan^{-1}\left[\frac{\psi}{\sin^2\psi + 1/kw}\right]$$

and if $w \gg w_m \cdot kw_m$ this becomes

$$\chi = \tan^{-1} kw_m.$$

w can be large in comparison with w_m, so that ψ can be neglected in the expression $(\psi + \chi)$. It is then found that as $(w - w_m)$ is increased $\sqrt{2\epsilon/m}\,\rho v$ rapidly tends to the constant values I_A and $(2I_T - I_A)$ respectively on the anode and screen sides of the potential minimum (see Fig. 3). The calculations then give

$$x_1/x_a = \frac{(1 + 2w_m)\,(1 - w_m)^{\frac{1}{2}}}{\sqrt{2J_T - J_A}}$$

$$x_2/x_a = \frac{(w_2 + 2w_m)\,(w_2 - w_m)^{\frac{1}{2}}}{\sqrt{J_A}}$$

so that

$$1 = \frac{(1 + 2w_m)\,(1 - w_m)^{\frac{1}{2}}}{\sqrt{2J_T - J_A}} + \frac{(w_2 + 2w_m)\,(w_2 - w_m)^{\frac{1}{2}}}{\sqrt{J_A}}$$

In these equations we may put

$$J_A = J_T\,(2/\pi)\tan^{-1} kw$$

and on plotting obtain the anode-current—anode-voltage characteristic as shown in Fig. 7. The region of negative slope extends from W_b back to W_K with an abrupt change at Wb intermediately. If W_2 lies between W_K and $W_{\sigma\sigma}$ there are seen to be *three* possible values of J_A for an assigned value of W_2. As k tends to infinity* the upper parts of the curve become coincident, so that the distinction is no longer graphically evident. The condition that $k = \infty$ represents the case in which a "virtual cathode" is formed between screen and the anode (when I_A is less than I_T). Fig. 7 shows that, as already stated, the corresponding "virtual cathode" knee-voltage, given by W_σ is extremely high in comparison with the practical values obtained by measurement, which are normally much

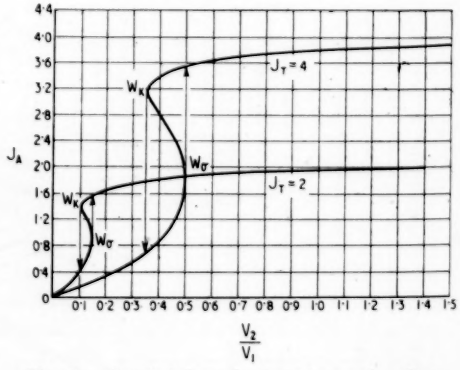

Fig. 6. *Computed anode-current—anode-voltage characteristics.*

more in agreement with the value of W_σ given by $k \simeq 5$.

* Or, if $I_A = I_T$ abruptly above some value of w_m, as in the distributions hitherto studied.

Acknowledgement

The writer wishes to accord his thanks to Mr. E. Y. Robinson, Chief Engineer of the Cosmos Manufacturing Co., Ltd., for his permission to publish this article.

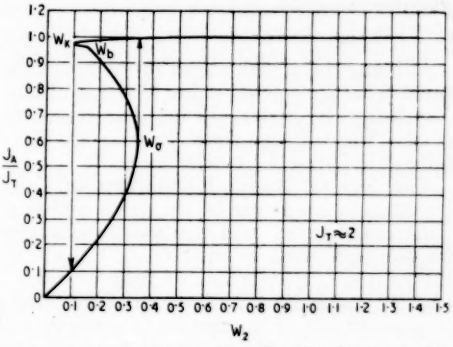

Fig. 7. *Anode-current—anode-volts characteristic when k is large.*

APPENDIX

Expressions (10) and (11) are derived from integrals of the form

$$\int_o^{w_m} \frac{i(s)\ ds}{\sqrt{w^2 - s^2}} \qquad \text{where } w > w_m$$

These occur, for example, in equations (7), (8), (9) and (10).

Let $i(s)$ be developed as a Fourier Integral for an even function.

$$i(s) = \int_o^{\infty} b(\lambda) \cos \lambda s \, d\lambda$$

where $$b(\lambda) = 1/\pi \int_{-\infty}^{+\infty} i(z) \cos \lambda z \, dz$$

Thus if $$i(s) = \frac{ck}{1 + k^2 s^2} \qquad b(\lambda) = c\epsilon^{-\lambda/k}$$

Substitute $$s = w \sin \phi$$

then $$\frac{b(\lambda) \cos \lambda s \, ds \, d\lambda}{\sqrt{w^2 - s^2}} = c \, d\lambda \, \epsilon^{-\lambda/k} \cos \left[\lambda w \sin \phi\right] d\phi$$

Now $$\cos (\lambda w \sin \phi) = J_o(\lambda w) + 2 \sum_1^{\infty} J_{2n}(\lambda w) \cos 2n\phi$$

where J_o J_{2n} are the Bessel Functions.

Integrate $\cos [\lambda w \sin \phi] d\phi$ from $\phi =$ zero up to $\phi = \sin^{-1} w_m/w$ i.e. up to ψ

$$\int_o^{\psi} \cos [\lambda w \, sm\phi] \, d\phi = J_0(\lambda w) \, . \, \psi + \sum_1^{\infty} J_{2n}(\lambda w) \frac{\sin 2n\psi}{n}$$

We next integrate

$$c\epsilon^{-\lambda/k} \left[J_o(\lambda w)\psi + \sum_1^{\infty} J_{2n}(\lambda w) \frac{\sin 2n\psi}{n} \right] d\lambda$$

from $\lambda = o$ to $\lambda = \infty$

using the standard result $$\int_o^{\infty} \epsilon^{-\lambda/k} J_{2n}(\lambda w) d\lambda = \frac{k}{\sqrt{1 + k^2 w^2}} \left[\frac{\sqrt{1 + k^2 w^2} - 1}{kw} \right]^{2n}$$

Then $$\int_o^{w_m} \frac{i(s)\ ds}{\sqrt{w^2 - s^2}} = ck \cos \alpha \left\{ \psi + \sum_1^{\infty} \tan^{2n} (\alpha/2) \frac{\sin 2n\psi}{n} \right\} = ck \cos \alpha \left\{ \psi + \chi \right\}$$

where

$$\tan \alpha = kw \, ; \, \chi = \tan^{-1} \left[\frac{\sin 2\psi \, . \, \tan^2 (\alpha/2)}{1 - \cos 2\psi \, . \, \tan^2 (\alpha/2)} \right]$$

c is chosen to meet the requirement that the total current $= I_T$, when Equ. (2) is integrated with limits for s equal to $+ 1$, $- 1$.

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[1] Rodda, *Wireless Engineer*, June 1936, p. 315.
[2] Strutt and van der Ziel, *Physica*, Oct. 1939, p. 977.
[3] Rodda, *Science Forum*, June 1943.
[4] Walker, *Wireless Engineer*, April, May, June 1945, pp. 157, 212, 276.
[5] Rodda, Lecture, Institute of Physics, Oct. 1944. *Electronic Engineering*, June, July, Aug., 1945.

RADIATION OF SHIP STATIONS ON 500 kc/s*

By J. Marique

(*Professor, Institute of Telecomunications, University of Brussels*)

Introduction

THE question of the radiation of ship stations on 500 kc/s is of more than purely academical interest, since the distress traffic takes place on this frequency and, in order that this traffic be effective, it is necessary that the range be sufficient. It is obvious that ships in a position to help should be within range under all conditions and that the distressed ship should receive their messages. The International Convention for Safety of Life at Sea[1] imposes minimum ranges to be obtained on 500 kc/s from ship stations transmitting on waves of types A2 and B, assuming that the receiving set makes use of a crystal. Nevertheless, as the range of a transmitter is not well defined without precise measurements, which are not normally carried out, the Convention has established, on the data they had at their disposal at the time (1929), the following rule :—

" Unless a more precise and practical method is available to determine the range of transmitters it is recommended that, as a guide, the following relations between the range in nautical miles (from ship to ship under normal conditions in daytime) and the power of the ship transmitter in metre-amperes for 500 kilocycles per second (600 m) be used :

100 nautical miles	..	60 *MA*
80 nautical miles	..	45 *MA*
50 nautical miles	..	25 *MA*

* MS. accepted by the Editor, November 1945.

M being the actual height in metres of the aerial from its highest point to the load line,

A being the current in amperes measured at the base of the aerial in case of B, or fully modulated A2 transmitters."

The General Radio Regulations of Cairo (1938) have made use of the same notion of " metre-amperes " and this value calculated for 500 kc/s must be inserted in the official list of the " Particulars of Ship stations " (G.R.R. appendix 8). The aim of this study is to make use of the results of a number of measurements made in actual conditions of working on type A2 waves, in order to establish a more rational basis of calculation for the predetermination of the ranges of ship stations. Most of these results were obtained before the war in experiments conducted by the International Radio-Maritime Committee.

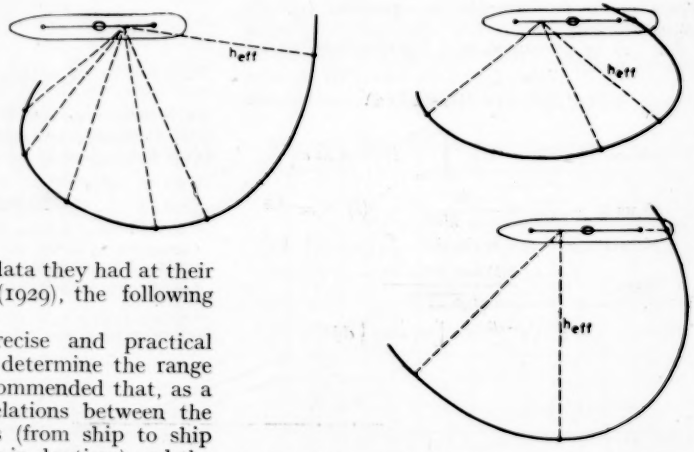

Fig. 1. *The apparent effective height of the aerials on three different ships is shown as a function of horizontal angle.*

1. Radiation Diagrams of Ship Aerials

It is well known that although ship aerials are generally of simple shapes, they are installed near numerous metal elements of large dimensions such as funnels, masts, rigging, etc., which influence the radiation. The radiation polar diagram must, therefore, be far from circular.

By measuring the field strength and the intensity of the current at the base of the aerial, it is possible to calculate the radiated power and the effective height of the aerial. Taking advantage of a bend of the river Scheldt where the course of ships changes by about 120° in a few minutes, it has been possible to carry out special measurements which made it possible to draw a part of the radiation diagram for three ships, as represented in Fig. 1. The full diagram is probably symmetrical to the fore and aft line. The frequency chosen was 454 kc/s (660 m) in order to avoid the jamming on 500 kc/s ; on the latter frequency the diagram must be very similar.

It can be seen that the apparent effective-height is not the same in all directions, and the graphs show that it can vary in the ratio 1 to 2 according to the direction in which the measurement is made. This well-known fact justifies the use of mean values calculated from several measurements when calculating ship-aerial radiation, and it will be adopted in the course of this study.

2. Radiated Power *v.* Metre-Amperes

In a paper published in 1939[2], we gave the results of numerous measurements of the power radiated by ship stations on 454 kc/s, the measurements being made at rather short distances. We showed that, in using the classical radiation formulae :

$$E = 120\pi \frac{\alpha h}{\lambda} \cdot \frac{I}{d} \cdot f(\rho) \quad .. \quad .. \quad (1)$$

$$W_r = 160\pi^2 \left(\frac{\alpha h}{\lambda}\right)^2 I^2 \quad .. \quad .. \quad (2)$$

and in considering the antenna height H as defined by the International Convention for Safety of Life at Sea, the following average relation between the radiated power W_r and the number of metre-amperes MA was found :

$$W_r = \left(\frac{MA}{K}\right)^2 \quad .. \quad .. \quad .. \quad (3)$$

K being a function of the form factor α of the antenna and of the wavelength :

$$K^2 = \frac{\lambda^2}{160\pi^2\alpha^2}$$

Fig. 2 shows the experimental results (small circles) and three curves corresponding to three values of K in equation (3). In log — log coordinates, these curves are straight lines. It must be borne in mind that each dot represents the mean value of several measurements made from the same ship, in order to eliminate as far as possible the influence of the radiation diagram.

Experience shows that K varies approximately between 25 and 50 at 454 kc/s. $K = 35$ seems to correspond the best to the greatest number of measurements made on this frequency.

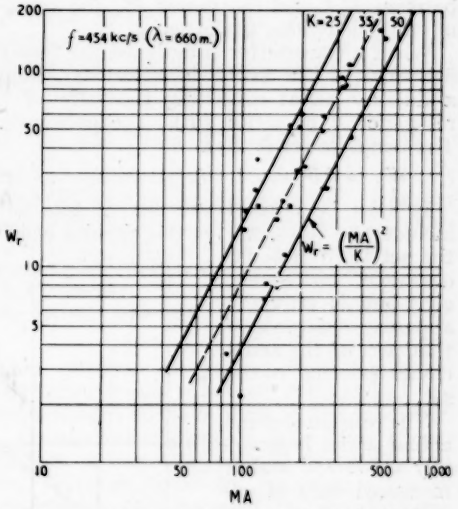

Fig. 2. *Measured results are shown by circles and the relation between radiated power and metre-amperes by the solid lines.*

From the values 25 to 50 it is deduced that the form factor lies between 0.33 and 0.66, *i.e.* that the effective height of ship aerials is between one-third and two-thirds of their actual height as defined by the Safety Convention.

As seen above, K is proportional to the wavelength :—for 500 kc/s, the extreme values of K are then 22.5 and 45 and the average value to be adopted for normal radiation conditions is $K = 32$.

The value of K does not seem to have any definite relation to the actual height of the aerial, since high values of K (which correspond to bad radiation) were experienced even with aerials as high as 40-50 metres.

From the above formulae it is possible to calculate the field strength corresponding to the values imposed by the Safety Convention for the ranges and numbers of metre-amperes. It is found that, in the

worst conditions ($K = 45$) the field strength at the limit of the range must be of the order of 50 to 60 μV/m.

3. Metre-Amperes *v.* Antenna Power

As the number of MA can be used to calculate from the above formulae the radiated power W_r and consequently the field strength at a distance, it is interesting to investigate if there exists a simple relation between MA and the power W_a the transmitter is delivering to the aerial.

When a transmitter is designed the power delivered to the aerial is measured, using a dummy aerial consisting generally of a resistance R in series with a capacitance. The aerial power is then

$$W_a = RI^2 \quad .. \quad .. \quad .. \quad .. \quad (4)$$

It must be pointed out that R does not include the resistance of the circuits inside the set itself (coil inductance, variometer, etc.) but is only the apparent resistance of that part of the aerial circuit external to the set.

The resistance of the actual aerial is generally unknown, and is measured only in exceptional cases; but it is known that its value lies normally between 4 and 8 ohms, and often 6 ohms is considered as an average value.

If the actual value of R were known, the measurement of I would permit the determination of W_a from equation (4), and of MA when the actual height H is known. In these circumstances the whole determination of the performance of a given ship station would be very easily dealt with.

Let us now investigate if this simple method, applicable to individual cases, could be generalized. At first sight, this may seem questionable, because ship aerials are so diverse, their heights are so different, their capacitances and natural wavelengths show such deviations that it would not be unsound to expect very important discrepancies which would prevent the use of a general method of calculation.

In order to check this fact, we have carefully examined the first-hand data concerning about 150 ship installations of A2 type transmitters of various nationalities. In

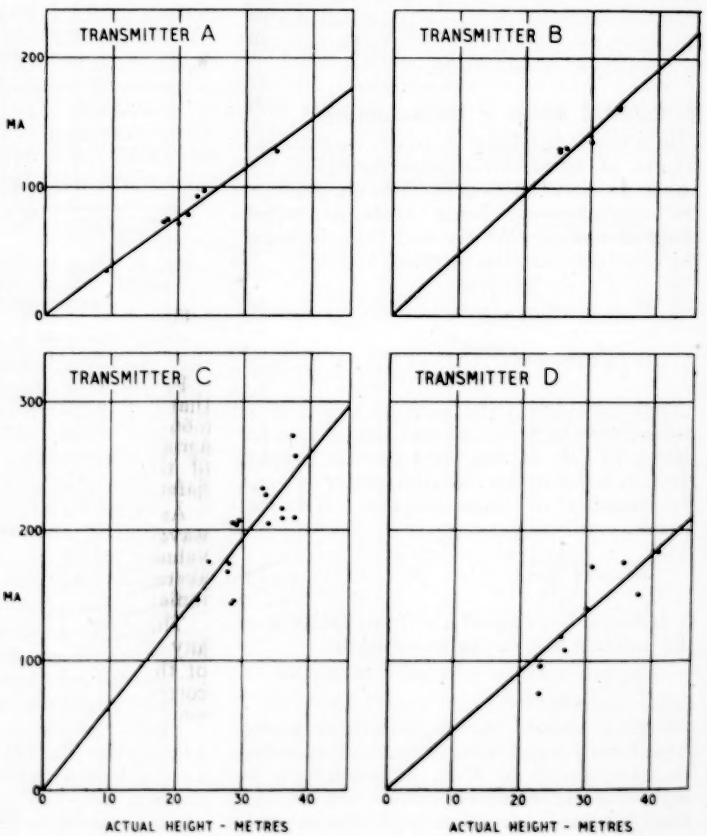

Fig. 3. *These graphs show measured values of MA for four different transmitters as a function of actual aerial height. Each dot corresponds to a ship.*

the graphs of Fig. 3 we have plotted the number of MA obtained as a function of the actual height for four different types of transmitters installed on several ships. Each dot corresponds to a ship. The number of MA was chosen as ordinate instead of the current I, since it is one of the figures to be published in the official lists of ship stations.

One can realize that in practice MA varies nearly proportionally to the actual height H, which means that the intensity of the current delivered by a certain type of transmitter is nearly constant whatever the ship aerial to which it is connected may be. It is then possible to write for each type of transmitter the average relation

$$I = \frac{MA}{H} = \text{constant} \quad .. \quad .. \quad (5)$$

In each of the four graphs of Fig. 3, the mean relation between MA and H can be represented by a straight line, the slope of which corresponds evidently to the mean value of the current. It must be pointed out that for the transmitter A, the proportionality between MA and H holds for aerials having actual heights between 9.2 m and 35 m.

This leads to the conclusion that it is by no means unreasonable to consider, for each type of transmitter, a mean value of the aerial current I characterizing that type of set. The discrepancies which may be experienced from one installation to another, and which, in the 150 data examined, do not exceed $\pm$ 33 per cent. of the mean values, are certainly due to local circumstances such as absorption by the rigging, very abnormal electrical characteristics of the aerial, or even to bad matching of the output circuit.

Incidentally, it is worth while to point out that manufacturers mention sometimes as "rated aerial power" the power which a transmitter is capable of delivering to the aerial, as defined in equation (4).

If, on the other hand, it has been possible to collect experimental data on actual installations of the same type of transmitter, and thus to calculate the corresponding mean value of the aerial current, a mean experimental value of the aerial resistance can be deduced by the reverse process, assuming that in each case the rated power has been actually delivered to the aerial

In doing so for four different types of transmitters, we obtained the figures of Table 1.

TABLE 1

Rated aerial power (Watts)	Experimental mean aerial current (Amps)	Deduced mean value of resistance (Ohms)
100	3.8	7
200	6.35	5
250	6.5	5.9
1000	12.2	6.8

These figures, however questionable the method of obtaining them may be considered, are well within the above-mentioned limits of 4 to 8 ohms.

4. Radiated Power W_r *v.* Aerial Power W_a

The considerations developed in Sect. 3 have shown that it is reasonable to consider a mean value of the aerial current characterizing each type of transmitter. The power delivered to the aerial is then

$$W_a = RI^2 = R\left(\frac{MA}{H}\right)^2 \quad .. \quad .. \quad (6)$$

This implies, of course, that the aerial resistance R has a normal value, which we shall assume to be 6 ohms in the following.

On the other hand, it has been shown in Sect. 2 that the radiated power W_r and the MA are fixed by equation (3). By eliminating MA between (3) and (6), one finds :

$$W_r = \frac{H^2}{R \cdot K^2} \cdot W_a \quad .. \quad .. \quad .. \quad (7)$$

The term $\frac{H^2}{RK^2}$ represents the ratio of the radiated power to the aerial power, or, in other words, the fraction of the aerial power that is radiated.

Assuming the average conditions :

$$R = 6 \text{ ohms} \qquad K = 32$$

it comes to : $\frac{W_r}{W_a} = \frac{H^2}{6000}$ (see Table 2)

These figures show, for instance, that under the average conditions of radiation ($K = 32$) and with an aerial of 6 ohms resistance, an aerial of 25 m actual height radiates only one-tenth of the power delivered by the transmitter.

Under the worst conditions of radiation ($K = 45$), $\frac{H^2}{RK^2} = \frac{H^2}{12000}$

TABLE 2

H	$\frac{W_r}{W_a} = \frac{H^2}{6000}$
10 m	0.0167
15	0.038
20	0.067
25	0.105
30	0.150
35	0.205
40	0.27
45	0.24
50	0.42

and the radiated power is only one-half of the figures of Table 2.

5. Generalized Graph of Ship Station Radiation

It is possible to put together all the above-mentioned elements and to draw a graph giving in a very simple way, under the average conditions met with in practice, the value of the field strength at any distance from a ship when the actual aerial height and the aerial power are known.

We found above the relation (7) which exists between the aerial power and the radiated power. Knowing the value of the latter, it is possible to draw the curve of variation of the field strength with distance. It is, therefore, only necessary to make use of the propagation curves published on the matter. The latest available are those calculated by the propagation of radio-electric waves sub-committee of the C.C.I.R. and presented to the Cairo Radiocommunications Conference by the British Administration[4].

In this document the curve of propagation during daytime over sea for 500 kc/s gives the value of the field strength as a function of the distance for 1 kW radiated. It is easy to deduce the curves for any given radiated power.

By a change of coordinates, the curves can be drawn to the axes " radiated power—distance " with the field strength values as parameter. In this manner the right-hand side part of Fig. 4 was obtained.

The curves of the above-mentioned sub-committee are drawn only for distances greater than about 130 km. For smaller distances, we calculated the field strength from the Sommerfeld-van der Pol formulae (1) and (2) mentioned at the beginning of this paper; this gives the dotted line curves of Fig. 4.

On the other hand, equation (7) can be represented by a set of curves to the axes ($W_r - H$) with W_a as parameter as drawn on the left-hand side of Fig. 4 for $R = 6$ ohms and $K = 32$.

The ordinate axis corresponds to the radiated power for both sides of Fig. 4, thus giving the connection between them.

The use of the graph is self-evident, the left-hand side elements being connected to the right-hand side ones by a simple construction like ABCDE drawn as an example.

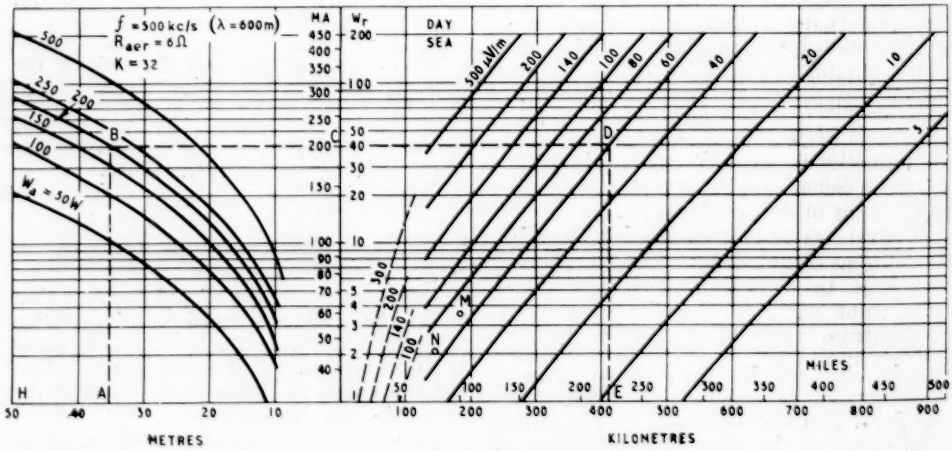

Fig. 4. *Generalized curves giving the relation between aerial power, aerial height and the field strength at the receivers for daytime conditions over sea on 500 kc/s.*

This construction shows that a transmitter delivering 200 watts in a ship aerial of 35 m actual height and 6 ohms resistance radiates under average conditions ($K = 32$) a power of 41 watts and gives during daytime over sea on 500 kc/s, a field strength of 60 μV/m at a distance of 415 km.

Under the worst conditions of radiation ($K = 45$) the radiated power would be one-half of the previous figure, *i.e.*, 20 watts and the same field strength figure would be obtained at 350 km.

Thanks to equation (3) it has been possible to graduate the ordinate axis in radiated watts and in metre-amperes assuming the value of K to be 32. The graph gives then at the same time the number of MA and the corresponding radiated power: the above-mentioned transmitter gives for instance 200 MA.

It must be pointed out that the right-hand side of the Fig. 4, representing the propagation curves of the wave of 500 kc/s during day-time over sea, is independent of the conditions of the aerial. On the contrary the curves of the left-hand side and the graduations in MA depend on these conditions. They could of course be calculated for any other value of R or K.

Fig. 4 can of course be used without change for $A1$ as well as $A2$ types of transmission, the values of the current intensity and of the field strength used above being all R.M.S. values. But it must be borne in mind that the power of a given transmitter is generally not the same for both types of waves.

Remark.—The points M and N on the right-hand side of Fig. 4 correspond to two figures ruled by the Safety Convention; M corresponds to 60 MA giving a range of 100 miles and N to 45 MA with a range of 80 miles. The corresponding values of the field strength are about 70μV/m for $K = 32$. The third figure (45 MA giving a range of 50 miles) cannot be plotted since it corresponds to a radiated power of less than 1 watt, which is the lowest limit of the graph.

6. Conclusion

Taking advantage of numerous experimental data obtained from actual installations on board ships, we have designed a generalized graph drawn for the average conditions met with in ship transmissions on 500 kc/s during day-time over sea: this graph gives the average relations between the actual height of the aerial, the power delivered by the transmitter, the MA, the radiated power, the field strength and the distance.

This result is based on the existence of a characteristic value of the aerial current for each type of ship transmitter or, in other words, on the approximate constancy of the ratio $\frac{MA}{H}$ for a type of transmitter whatever the height of the aerial may be.

This implies of course that the transmitter is effectively able to deliver its rated power to any type of ship aerial, *i.e.*, that it is provided with the necessary elements of electrical matching.

REFERENCES

[1] "International Convention for the Safety of Life at Sea," London, 1929. Art. 31. §§8, 10 and 12.

[2] "Relation between the Power Radiated by a Ship Station and the Number of 'Metre-Amperes.' *L'Onde électrique*, Feb. 1939, pp. 92–96. Summarised in *The Wireless Engineer*—June 1939, p. 321.—In this abstract, the formula should be read $W = (MA/K)^2$, instead of $W = MA/K$.

[3] van der Pol. Z.f.H. April 1931.

[4] "Conference Internationale des Radiocommunications," Cairo 1938. Vol. I—Doct. 558R; p. 445.

University of Glasgow

DR. BERNARD HAGUE has been appointed to the James Watt Chair of Electrical Engineering at the University of Glasgow which is being vacated later in the year by Dr. G. W. O. Howe. Born in 1893, Dr. Hague was educated at the Grammar School, Eccles and the Central School, Rochdale. After practical training with Carter Bros., Rochdale, and Ferranti, Ltd., he proceeded to the Department of Electrical Engineering of the City and Guilds College. He graduated B.Sc. (London) and received the diploma of the Imperial College for advanced study and research. In 1919 he proceeded to the London M.Sc., and in 1927 to the D.Sc. degrees, while he took the Ph.D. (Glasgow) degree in 1926.

From 1916–1920 Dr. Hague was associated with the Royal Aircraft Establishment, afterwards becoming a lecturer at the City and Guilds College. Since 1923 he has occupied positions at the University of Glasgow, being given leave of absence in 1929–30 to act as visiting professor in the Polytechnic Institute of Brooklyn, New York. He has published four books and twenty-one papers on electrical and allied subjects.

Electronic Semi-Conductors

A LECTURE on "Electronic Semi-conductors" will be given by Dr. R. W. Sillars at a joint meeting of the Institute of Physics and the Institution of Electronics to be held at 7 p.m. on Friday, May 17th, in the large physics lecture theatre at the University, Manchester. Tickets may be obtained from Dr. F. A. Vick at the University, Manchester, or from L. F. Berry, 105 Birch Avenue, Chadderton, Lancs.

CORRESPONDENCE

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

"Time-base Converter and Frequency-divider"

To the Editor, "Wireless Engineer."

SIR,—In connection with the discussion on the respective merits of the above and of the "signal converter," I am gently chided by Messrs. Nagy and Goddard in your October 1945, issue for what they are pleased to describe as an unprogressive attitude. It will be recollected that in my original article in the August 1945 issue, I suggested that it was a retrograde step to employ a new and specialized type of valve (the signal converter) for an application which could be met by the use of conventional valves.

It seems necessary for me to expand my thesis a little more fully, although it must, of course, be admitted that discussions along these lines are matters of opinion and not of fact.

My contention is that it is undesirable to produce a very large number of *types* of components and that the introduction of any new type is unjustified unless it leads to technical or economic superiority of some finished instrument as compared with what can be achieved by the use of existing components. It will be noted that it is not a sufficient condition for the new component to replace several existing ones, since the cost of components depends enormously on their demand and the process of rigorous limitation of component types along the lines discussed would greatly reduce production costs. To me, an examination of any valve data manual is a somewhat depressing reflection on the lack of collaboration which exists between designers, and I view with alarm the suggestion that deflection-modulated valves should be added to the huge list of types now available.

Wave of 1.2 Mc/s taken with a time-base frequency of 200 kc/s.

Ultimately, the acid test of justification is the relative performance and cost of the finished instrument. At the moment it does not seem to me that Messrs. Nagy and Goddard have succeeded in establishing any great superiority in time-base technique as a result of the use of the signal converter. In addition, their letter in the October issue contains one or two statements of a very controversial nature.

Speaking about driven time bases in general, and my own in particular, they say, "When the frequency is such that the capacitances are of the order of the resistances, synchronization is unreliable and the linearity of the scan is destroyed." This observation seems inaccurate, because a driven time base is not synchronized, and the only effect of raising the frequency is to reduce progressively the steepness of the wave fronts of the discharge pulses from the squaring circuits. As a result, the commencement of the flyback time becomes less definite. However, the stability of the pattern can be in no way affected, since the device is working under cyclic and repetitive conditions. Included herewith is a photo taken on a driven-type time base at a sweep frequency of 200 kc/s (1.2 Mc/s wave). The exposure time was 20 seconds. The sharpness of the flyback shows the high stability obtainable.

Later on they suggest that the signal converter is superior to the conventional valve because two parameters are available in its operation. Presumably they imply (correctly) that this leads to greater range of application. However, I contend that this is a doubtful advantage, for looked at from another angle it also implies that closer control of the device is necessary to secure a given result. Surely one of the most desirable developments in circuit technique generally is the construction of networks which will do one thing only and which do not require close control of operating voltages, etc., in order to keep their performance on the "straight and narrow path."

However, the major drawback of the signal converter appears to me to reside in its high output impedance and in the necessity for a load which is almost wholly reactive. Appendix 1 in their September article indicates the result of a shunt resistance across the charging capacitor. It appears that a shunt resistance of 10-megohms value results in a 20 per cent. linearity error. Reference to my article on deflector-plate characteristics of cathode-ray tubes in the *Wireless Engineer* for August 1941 will show that for positive deflector-plate excursions 10 megohms is an optimistic estimate for the shunt resistance on an average tube and, therefore, it would appear that a high degree of linearity cannot be expected. In addition, it is an axiom of cathode-ray tube work that the deflector-plate driving-source impedance should be kept low and this condition is violated by the "signal converter."

The trace photographs published by Messrs. Nagy and Goddard indicate a much higher degree of linearity than would be expected from the theoretical reasoning, unless the time-base plates were pre-biased so as to operate always negatively with regard to the final anode. An examination of the circuit diagrams indicates that no special steps have been taken to ensure such negative operation under all conditions, and I should be glad if they could suggest an explanation for the discrepancy.

HILARY MOSS.

Highbury, N.5.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

573 857.—Push-pull amplifier, particularly for indicating physiological reactions, in which provision is made for offsetting an input circuit which is not balanced to earth.

Standard Telephones and Cables Ltd. (assignees of J. R. Banker). Convention date (U.S.A.) 23rd December, 1942.

573 869.—Telephone or like relay for regenerating and responding to definitely-spaced sequences of signal-impulses of uncertain duration.

Standard Telephones and Cables Ltd., G. C. Hartley, and W. J. Reynolds. Application date 15th December, 1942.

AERIALS AND AERIAL SYSTEMS

573 762.—Flexible mounting for the rod aerial of a motor vehicle, to prevent breakage by impact with the branch of a tree or similar obstruction.

D. W. Mitchell. Application date 27th November, 1942.

574 418.—Aircraft aerial made of a rod of plastic material which is metallically coated for part of its length, the rod being supported where it is uncoated and therefore non-conducting.

Boulton Paul Aircraft Ltd. and J. D. North. Application date 29th November, 1940.

DIRECTIONAL WIRELESS

573 774.—Remote-control tuning arrangement, particularly suitable for a multi-band automatic direction finder.

A. H. Stevens (communicated by Lear Avia Inc.). Application date 24th November, 1943.

573 896.—Spiral aerial of controllable directivity, wherein the length of each turn, and the spacing between turns, are specifically related to the operative wavelength.

Marconi's W.T. Co. Ltd. and C. S. Franklin. Application date 24th March, 1941.

573 922.—Offsetting the effects of banking and lateral drift when an aeroplane is homing on to a radio transmitter.

S. Smith & Sons (Motor Accessories) Ltd. and F. W. Meredith. Application date 6th April, 1940.

RECEIVING CIRCUITS AND APPARATUS

(See also under Television)

573 168.—Receiver, including a frequency-doubling device, for restoring the higher harmonics that are normally lost owing to circuit, or other, limitations.

Ferranti Ltd. and M. K. Taylor. Application date, 2nd December, 1943.

573 354.—Shunt circuit including a non-linear device, such as a copper-oxide rectifier, for automatically allowing the signal current to by-pass an amplifier or repeater that has failed.

Automatic Telephone and Electric Co. Ltd. Convention date (U.S.A.) 5th March, 1943.

573 837.—Coaxial-line end-coupling and mounting for a crystal detector of the "cat's-whisker" type, designed for rectifying centimetre waves.

The British Thomson-Houston Co. Ltd., T. H. Kinman, and B. A. C. Tucker. Application date 1st May, 1944.

573 978.—Reducing strong interference by superposing locally-generated oscillations so as to form an interference field-pattern having a voltage-node at the pick-up point.

J. Robinson. Application dates 4th April, 1940, and 29th April, 1941.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

573 272.—Television scanning system in which the higher harmonics of the modulated scanning frequencies are eliminated, in order to correct for "keystone" distortion.

Farnsworth Television and Radio Corporation. Convention date (U.S.A.) 1st April, 1942.

573 274.—Method of modulating the scanning frequencies, and for cutting-out the resulting harmonics, in order to correct "keystone" distortion in television.

Farnsworth Television and Radio Corporation. Convention date (U.S.A.) 30th April, 1942.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

573 269.—Controlling the periodicity of amplitude changes in a thermionic or like oscillator by utilizing the transit-time of a group of pressure waves in passing through a liquid.

Scophony Ltd. and S. H. M. Dodington. Application date 19th September, 1940.

573 406.—Frequency-modulation system in which a carrier-wave is combined with the current produced from a photo-electric cell under the action of a beam of light, after it has been passed through crossed Nicol prisms and a Kerr cell to which the signal voltage is applied.

Philips Lamps Ltd. Convention date (U.S.A.), 11th July, 1942.

573 508.—Heptode valve with two discharge paths in series, one of which operates as a cathode-follower, the combination serving as a multivibrator to generate either continuous oscillations, or a series of pulses.

T. E. Ivall. Application date 19th August, 1943.

573 629.—Secret telephony system in which control signals, based on the vowel frequencies, and otherwise, are incorporated to facilitate accurate restora-

tion of the original message at the receiving end.

" Patelhold " Patentverwertung & Holding A.G. Convention date (Switzerland) 3rd December, 1941.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

573 207.—Short-wave generating-valve, of the beam-deflection type, with auxiliary electrodes to allow the electron-stream to be decelerated cyclically whilst maintaining full output.

J. H. O. Harries. Application date 5th September, 1940.

573 208.—Mounting and cooling arrangement for a short-wave generating-valve, of the beam-deflection type, connected to coaxial-line input and output circuits.

J. H. O. Harries. Application date 5th September, 1940.

573 209.—Method of ensuring accurate alignment of the electrodes of a high-power valve of the beam-deflection type.

J. H. O. Harries. Application date 23rd September, 1940.

573 232.—Gas-filled discharge-tube fitted with a resilient screen, in order to control and localize the ionization, and to minimize ionic bombardment of the cathode.

Western Electric Co. Inc. Convention date (U.S.A.) 5th June, 1942.

573 395.—Location of an auxiliary grid for controlling the electron velocity and concentration in a valve of the beam-deflection type (addition to 571 678).

G. R. E. Cleveland. Application date 27th July, 1944.

573 569.—Arrangement and spacing of the electrodes of a diode or other double-ended discharge tube, particularly of the midget type, for short-wave working.

Ferranti Ltd., A. L. Chilcot, and S. Jackson. Application date 18th April, 1942.

573 570.—Process for coating the cathode of a valve, after evacuation of the bulb and degassing of the electrodes.

Ferranti Ltd. and A. L. Chilcot. Application date 18th April, 1942.

573 571.—Jig process for assembling and spacing the electrodes of a short-wave or like valve, where the input and output leads are located at different parts of the bulb.

Ferranti Ltd., A. L. Chilcot, and J. L. Miller. Application date 18th April, 1942.

573 572.—Jig process for the assembly of a diode or like discharge tube comprising two cup-like parts.

Ferranti Ltd., A. L. Chilcot, S. Jackson, and F. W. Taylor. Application date 18th April, 1942.

573 587.—Discharge tube of the cathode-ray type in which a high-frequency transformer is housed inside the evacuated envelope.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) 7th December, 1942.

573 605.—Multi-pin valve-socket in which one pin is mounted, asymmetrically with the others, on a metal band surrounding the socket.

Standard Telephones and Cables Ltd. and W. T. Gibson. Application date 2nd November, 1943.

573 612.—Process for coating the cathode of a glow-discharge tube with an alkali metal, to reduce the striking voltage.

Standard Telephones and Cables Ltd. and R. R. Back. Application date 3rd December, 1943.

573 852.—Arrangement for controlling the " gettering " process, in order to prevent undesirable deposition, particularly in the manufacture of high-powered diode rectifiers.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 13th October, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

573 173.—Photo-electric cell fitted with retarding and collecting electrodes to ensure a response to illumination over a large solid angle.

W. Blackman, Cinema-Television, Ltd. and G. A. R. Tomes. Application date 13th December, 1940.

573 345.—Bridge-circuit input for cyclically displacing the ignition-point of a grid-controlled discharge-tube so as to stabilize the operation, say of a magnetron valve controlled by it.

Akt Brown, Boverie et Cie. Convention date (Switzerland) 23rd May, 1942.

573 348.—Inter-electrode resistance-coupling to enable a pentode or tetrode valve to be operated efficiently as a triode or diode, say for a voltage-stabilizer or full-wave rectifier.

B. M. Hadfield. Application date 5th November, 1943.

573 365.—R.F. transformer for centimetre waves, consisting of a hollow tuned resonator, into which are inserted, to a variable extent, the ends of the primary and secondary coaxial-line circuits.

J. Collard. Application date 3rd June, 1941.

573 405.—Selective remote-control relay in which a pair of valves with different working characteristics is arranged to give a " peak " output for a predetermined control-bias.

Standard Telephones and Cables Ltd. (assignees of G. Deakin). Convention date (U.S.A.) 14th September, 1942.

573 418.—Bridge circuit in which a pair of capacitors is varied, whilst maintaining a constant ratio, in order to measure inductance and " Q "

The General Electric Co. Ltd., H. C. Turner, and W. H. Fisher. Application date 24th September, 1943.

573 451.—Compact arrangement of an artificial line for attenuating ultra-high frequencies.

Standard Telephones and Cables Ltd. (assignees of C. B. Watts Junr). Convention date (U.S.A.) 17th August, 1942.

573 483.—High-frequency ignition system designed to off-set the variable-shunt effect due to the partial sooting of a sparking-plug.

The British Thomson-Houston Co. Ltd. and D. F. Welch. Application date 11th June, 1943.

573 621.—Projectile fitted with a fuse which is subject to remote radio-control so that it is detonated automatically in the proximity of the target.

G. W. Walton. Application date 11th January, 1941.